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THE CAUSES OF VOLTAGE AND CURRENT FLUCTUATIONS

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There are in principle only two causes for the occurrence of spontaneous current and voltage fluctuations in amplifiers: the thermal agitation of the electric charge in conductors and the motion of free electrons in amplifier valves. These "sources of noise" in amplifiers, which have been repeatedly discussed in this periodical, are here dealt with once again on the basis of new information which has lately been gained. Special emphasis is laid on the significance of cosmic phenomena in the noise on aerials and on the disturbing effects of the induced currents which are excited by the fluctuations in the emission of the cathode on the control grid of an amplifier valve.

The weak spontaneous current and voltage fluctuations which occur in every electrical circuit have repeatedly been discussed in this periodical [1]). These current and voltage fluctuations may give rise to undesired phenomena when they occur in an apparatus or a part of an apparatus in which extremely small currents and voltages are used, such as a microphone, a photocell or the aerial and the input circuit of a radio receiving set. In the last case the effect of the spontaneous current and voltage fluctuations is universally familiar; they lead to a continuous sound which is usually called the noise of the receiver.

As causes of the noise in receivers, in the articles referred to above, the resistances and the amplifier valves present in the circuit were mentioned. During the three years since the publication of these articles new information has been gained about the spontaneous current and voltage fluctuations in such circuit elements, which information is found to be of special importance in the case of reception on short waves. Since the combatting of the characteristic noise of the set is found to form one of the chief problems in the construction of short wave receivers, we shall again discuss the voltage and current fluctuations in resistances and electronic valves on the basis of the new information, while the results produced by the fluctuations on the performance of the receiving sets will be dealt with in a following article in this periodical.

1) M. Ziegler. The causes of noise in amplifiers, Philips techn. Rev. **2**, 136, 1937; Noise in amplifiers contributed by the valves, Philips techn. Rev. **2**, 329, 1937; Noise in receiving sets, Philips techn. Rev. **3**, 189, 1938.

Voltage and current fluctuations in resistances

The voltage and current fluctuations which occur in resistances may be considered as a thermal phenomenon: the more or less mobile electrical charge which is situated in the interior of a resistance is not only brought into motion by an applied voltage, but it also takes part in the thermal molecular motion, just as in the case of microscopically small particles suspended in a liquid (Brownian movement). This motion of the charge is manifested as an irregularly varying current of which it may be proved that in intensity and character it is entirely determined by the magnitude of the resistance and the temperature, while the nature of the resistance (for instance, whether carbon or wire resistance) is without effect [2]).

If the spontaneous current fluctuations are resolved into components with different frequencies, a continuous spectrum is found in which all frequencies occur in equal intensity, *i.e.* each frequency band Δf furnishes the same contribution to the effective current, independent of the frequency f itself. The mean square value of this current contribution is given by the formula

$$\overline{i_R^2} = \frac{4\,kT}{R}\,\Delta f, \quad \quad (1)$$

2) The thermodynamic considerations amount to the following: Two resistances A and B which have the same temperature are considered to be connected in parallel: the voltage fluctuations generated by A must then heat resistance B by the same amount as the fluctuations generated by B heat A. Otherwise there would be a temperature difference which would be contrary to the fundamental laws of thermodynamics. See on this subject also the first article referred to in footnote 1).

where $k = 1.38 \times 10^{-23}$ watt sec/degree represents the Boltzmann constant, T the absolute temperature (°K), R the resistance in ohms and f the frequency in c/s. The formula determines the fluctuation current in amperes which occurs when the resistance is short circuited. The fluctuation voltage v_R between the ends of an open resistance is often desired. This is connected with i_R by $v_R = Ri_R$, and therefore

$$\overline{v_R^2} = 4\,kTR\,\Delta f \quad . \quad . \quad . \quad . \quad . \quad (1a)$$

In order to give some idea of the order of magnitude of the fluctuations let us consider a resistance of 10^5 ohms. At room temperature and with a frequency band of 10^4 c/s, equation (1) gives an effective current of

$$\sqrt{\overline{i_R^2}} = 4 \cdot 10^{-11}\ \mathrm{A},$$

while from equation (1a) the following voltage fluctuations result:

$$\sqrt{\overline{v_R^2}} = 4\ \mu V.$$

Up to this point the fluctuation phenomena were also discussed in the article referred to above on the causes of noise in amplifiers. If we now apply the results to actual connections, various questions arise whose answers are not immediately obvious. The first question is the following. One of the resistances which may cause considerable disturbance by their voltage fluctuations is the resistance of the tuned oscillator circuit which is connected between control grid and cathode of the amplifier valve (see *fig. 1*). When we wish to calculate the voltage fluctuations on the grid, what value of the resistance must we choose, the resistance of the coil (r), which may be considered as the actual source of the fluctuations, or the resonance resistance R_{LC} of the whole LC circuit (L/Cr) which is connected between control grid and cathode?

The answer is that this must be a matter of indifference when the calculation is correctly performed. Since the voltage fluctuations over a

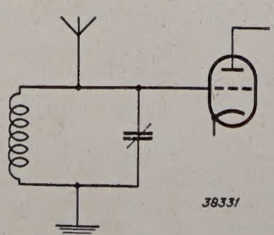

Fig. 1. Diagram of the connections of a high-frequency amplifier stage. Between cathode and control grid a high-frequency oscillator circuit is connected which may be considered as a resistance R_{LC} for the resonance frequency.

resistance of a given size are independent of the nature of the resistance, the voltage fluctuations over the oscillator circuit are determined by the resonance resistance R_{LC} in the neighbourhood of the resonance frequency. The mean square of the fluctuation voltage must therefore have the value

$$\overline{V_C^2} = 4\,kTR_{LC}\,\Delta f \quad . \quad . \quad . \quad . \quad . \quad (2)$$

The first mentioned argument is, however, also correct; the source of the fluctuations is exclusively the series resistance r of the coil, see *fig. 2* (other sources of damping, such as that of the condenser and the input damping of the valve will here be neglected), and when we assume a voltage fluctuation of

$$\overline{v^2} = 4\,kT\,r\,\Delta f$$

in series with the resistance r, we must also obtain the correct result.

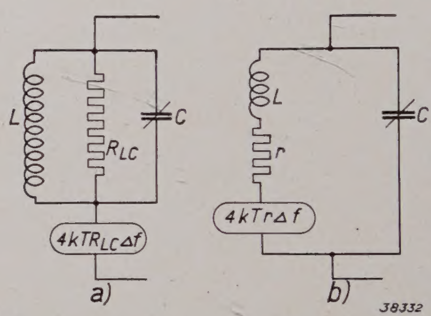

Fig. 2. *a*) Voltage fluctuations in series with the resistance R_{LC} of the oscillator circuit. *b*) Voltage fluctuations in series with the resistance r of the coil. If the damping of the oscillator circuit is caused only by the resistance r of the coil, the two diagrams are equivalent to each other.

The fact that the two lines of reasoning are actually equivalent is confirmed by simply calculating the voltage on the oscillator circuit, beginning with the fluctuation voltage over the resistance r. For this purpose we consider a component of the fluctuations having a given angular frequency, ω. The voltage fluctuations v then cause a current through the oscillator circuit which is given by

$$i = \frac{v}{r + j\left(\omega L - \dfrac{1}{\omega C}\right)}.$$

This current causes between control grid and cathode of the amplifier valve an A.C. voltage:

$$V_C = \frac{i}{j\omega C} = \frac{v}{(1 - \omega^2 LC) + j\omega rC}$$

and by taking the mean square over the time the

following result is obtained:

$$\overline{V_C^2} = 4\,kT\,\Delta f \frac{r}{(1-\omega^2LC)^2 + \omega^2r^2C^2} \cdot \quad (3)$$

In the special case of the resonance frequency, $\omega^2LC = 1$, and therefore

$$\overline{V_C^2} = 4\,kT\,\Delta f \cdot \left(\frac{L}{Cr}\right) \cdot$$

The expression in parenthesis is the resonance resistance R_{LC} of the oscillator circuit, so that equation (2) is confirmed.

In a frequency region outside the resonance frequency the following formula is valid for the impedance in parallel with the oscillator circuit:

$$Z_{LC} = \frac{r + j\omega L\left(1 - \omega^2LC - \dfrac{r^2C}{L}\right)}{(1-\omega^2LC)^2 + \omega^2r^2C^2} = X + jY.$$

The real part of this corresponds to the fraction in equation (3). From this it follows that the voltage fluctuations on an oscillator circuit for any given frequency can be described by the general formula:

$$\overline{V_C^2} = 4\,kT\,X\,\Delta f, \quad . \; . \; . \; . \; . \quad (4)$$

where X is the real part of the circuit impedance.

In order to be able to use equation (4) the frequency interval Δf must be chosen so small that within this interval the circuit impedance does not change to any extent. If one is concerned with a finite frequency interval, $X\Delta f$ must be replaced by $\int X df$. The exact expression for the mean square of the fluctuation voltage then becomes:

$$\overline{V_C^2} = 4kT \int_{f_1}^{f_2} X df = 4kT \frac{r}{2\pi} \int_{\omega_1}^{\omega_2} \frac{d\omega}{(1-\omega^2LC)^2 + \omega^2r^2C^2} \cdot$$

By the introduction of the resonance frequency $1/\sqrt{LC} = \omega_0$ and of the symbols $\omega/\omega_0 = x$ and $\omega^2r^2C^2 = a^2$ this formula can be written:

$$\overline{V_C^2} = \frac{4kT}{C}\,\frac{a}{2\pi} \int_{x_1}^{x_2} \frac{dx}{(1-x^2)^2 + a^2x^2} \quad . \; . \; . \quad (4a)$$

The integration is somewhat complicated, but does not offer any fundamental difficulties. The general integral is:

$$\frac{4kT}{C} \cdot \frac{a}{2\pi} \left[\frac{1}{2a} \operatorname{arctg} \frac{ax}{1-x^2} + \frac{1}{4\sqrt{4-a^2}} \ln \frac{x^2 + x\sqrt{4-a^2} + 1}{x^2 - x\sqrt{4-a^2} + 1}\right]_{x_1}^{x_2}.$$

We shall now consider two special cases more closely. Firstly the case, which is interesting from a physical point of view, where the integral extends over all frequencies ($x_1 = 0$, $x_2 = \infty$), secondly the case which is more important technically where only a certain region on either side of the resonance frequency furnishes a contribution.

In the first case the inverse tangent occurring within the bracket is found to possess a value π, while the second term between the brackets furnishes no contribution at the limits 0 and ∞. The result therefore is that

$$\overline{V_C^2} = \frac{4kT}{C}\,\frac{a}{2\pi}\,\frac{1}{2a} \cdot \pi = \frac{kT}{C} \quad . \; . \; . \; . \; . \quad (4b)$$

The total fluctuation voltage on the condenser is thus independent of the self-induction and the resistance in the oscillator circuit.

The result could indeed have been derived directly by a thermodynamic method. According to the law of equipartition, the following holds for the potential energy of the charged condenser:

$$\tfrac{1}{2}C\,\overline{V_C^2} = \tfrac{1}{2}kT,$$

from which (4b) follows directly.

We shall now pass on to the case where only a limited frequency region on either side of the resonance frequency furnishes a contribution to the fluctuation voltage. The width of this region is in general determined by the selectivety of the following stages, and may for example be chosen such that the resonance resistance has fallen to one half at the limits of the region.

From equation (4a) it may be seen that for the limiting frequencies x_1 and x_2 the absolute value of $1-x^2$ is then equal to ax. The argument of the inverse tangent in the general integral then has the value -1 at the lower limit and the value $+1$ at the upper limit. The inverse tangent thus covers a range of angles of 90° or $\pi/2$, namely from $-45°$ to $+45°$. The second term between the brackets, in the important case of a sharp resonance curve ($a \ll 1$), is again found to furnish no contribution. The result is therefore:

$$\overline{V_C^2} = \frac{4kT}{C} \cdot \frac{a}{2\pi} \cdot \frac{1}{2a} \cdot \frac{\pi}{2} = \frac{kT}{2C} \quad . \; . \; . \; . \quad (4c)$$

The fluctuation energy in the frequency region chosen is therefore one half of the total fluctuation energy.

For practical applications it is advantageous to write equation (4c) in a form which more nearly resembles equation (4), namely:

$$\overline{V_C^2} = 4kT\,\overline{X}\,\Delta f, \quad . \; . \; . \; . \; . \; . \; . \quad (4d)$$

where $\overline{X}$ is the average value of the real part of the circuit impedance in the frequency region in question. By the use of the integral given in the foregoing this average value can easily be calculated and one obtains:

$$\overline{X} = \frac{\pi}{4}\left(\frac{L}{Cr}\right) = 0{,}78\,R_{LC},$$

while for the frequency region at whose limits the real part of the circuit impedance has fallen to one half, a width of

$$\Delta f = \frac{1}{2\pi}\,\frac{r}{L}$$

is calculated. By substituting these values of $\overline{X}$ and Δf in equation (4d), formula (4c) should be obtained. One actually finds that:

$$\overline{V_C^2} = 4kT\left(\frac{\pi}{4}\,\frac{L}{Cr}\right) \cdot \left(\frac{1}{2\pi}\,\frac{r}{L}\right) = \frac{kT}{2C}.$$

If the sources of damping do not lie in the self-induction coil, but in the condenser, fundamentally the same results are found. For dampings which are caused by the amplifier valve, however, different results may be expected, about which some remarks will be made later on in this article.

Current fluctuations in the aerial

An oscillator circuit of special interest to us is the aerial. It has a definite impedance depending upon the frequency, the real part of which is usually called the radiation resistance of the aerial for the frequency in question.

This radiation resistance, according to the general law (1a) will also have to exhibit certain thermal fluctuations. This involves the question as to what its temperature is. It is clear that the temperature of the aerial wire has nothing to do with the fluctuations in question. The radiation resistance R_{ant} of the aerial means physically that when an A.C. voltage V_{ant} is applied to the aerial it dissipates an energy of:

$$P = \frac{\overline{V_{ant}^2}}{R_{ant}}$$

which is radiated into space. Conversely, the thermal voltage fluctuations must have their source in the fact that space is filled with thermal radiation (heat radiation) which is taken up by the aerial. If it is assumed that there is in space an equilibrium between the radiation and matter at a given temperature T, then the intensity and special distribution of the radiation is known. It is therefore possible to calculate the voltage fluctuations on the aerial caused by the radiation, and one finds, as was to be expected, that between the radiation resistance and the voltage fluctuations of an aerial there is is the general relation:

$$\overline{V_{ant}^2} = 4\,kT\,R_{ant}\,\Delta f$$

How high is the effective temperature T of space for heat radiation of radio frequencies? The voltage fluctuations on an aerial make it possible to give an experimental answer to this question. Experiments in this direction carried out by Jansky and Reber [3]) have led to very remarkable results. It is found that the radiation in the region of the high radio frequencies (metre waves) is mainly of cosmic origin and appears to come from the milky way. Instead of a temperature of several degrees absolute, as might be expected for the universe, a temperature of about 10 000 °K is found.

[3]) K. G. Jansky, A note on the source of interstellar interference, Proc. Inst. Rad. Eng. **23**, 1158, 1935. G. Reber, Cosmic static, Proc. Inst. Rad. Eng. **28**, 68, 1940; Astrophys. J. **91**, 621, 1940. For theoretical considerations of this radiation see: H. A. Kramers, Theory of the continuous X-ray spectrum, Phil. Mag. **46**, 836, 1932; A. S. Eddington, Diffuse matter in interstellar space, Proc. Roy. Soc. **111**, 424, 1926. J. A. Gaunt, Continuous absorption, Trans. Roy. Soc. **229**, 163, 1930. L. C. Henyey and P. C. Keenan, Interstellar radiation from free electrons and hydrogen atoms, Astrophys. J. **91**, 625, 1940.

On the basis of theoretical investigations [3]) it may be accepted as practically certain that the radiation which causes the noise of the aerial has its origin in interstellar matter which fills the milky way system with a concentration by weight corresponding to approximately one hydrogen atom per cm^3. The thickness of the layer is about 50 000 light years. Under the action of starlight this matter is ionized for the main part into electrons and positively charged particles. The irregular motion of the electrons, as a result of their mutual repulsion and their attraction of the positive particles, leads to a radiation which is intense enough to make the observed phenomena understandable in principle. A complete explanation will, however, only be possible when more data are available about interstellar matter.

Voltage and current fluctuations in amplifier valves

The currents in an amplifier valve have their source in the thermal emission of a heated cathode. If this process is considered fluctuation phenomena are immediately expected because of the fact that the negative charge emitted consists of electrons, *i.e.* particles, and is therefore emitted in definite multiples of $e = 1.6 \times 10^{-19}$ coulombs.

The number of electrons which is emitted by the surface of the cathode in a time interval of a given length will not always be exactly the same, but will exhibit accidental variations around a definite average value, just as, when it is raining, the number of drops which strike a definite spot on the ground is not always exactly the same for every second. When it is assumed that all the electrons emitted reach the anode the variations experienced by the anode current due to these fluctuations can easily be calculated. For the performance of this calculation we may again refer to the article cited in footnote [1]) on the causes of noise in amplifiers. The calculation leads to the result that the current variations, like the thermal current fluctuations in a resistance, cover a continuous spectrum with constant amplitude. The mean square of the current fluctuations i_k is given by the formula:

$$\overline{i_k^2} = 2\,e\,I_k\,\Delta f, \quad \ldots \ldots \quad (5)$$

where I_k represents the average emission current of the cathode [4]).

With the practically occurring values of I_k = 30 mA, Δf = 10 000 c/s, for example, one obtains

[4]) This result is valid only for frequencies at which the time of oscillation is still several times greater than the transit time of the electrons between cathode and anode. On the subject of the current fluctuations at still higher frequencies see S. Ballantine, J. Franklin Inst. **206**, 159, 1928.

$\sqrt{\overline{i_k^2}} = 10^{-8}$ A. This current fluctuation caused by the corpuscular nature of the electric charge, is usually called the shot effect of the amplifier valves.

By measurements with a diode the formula for the shot effect may be tested. If the anode voltage is chosen sufficiently high, the condition is indeed fulfilled that each electron which leaves the cathode reaches the anode. The diode is then said to be in the saturated state, since upon further increase of the anode voltage the anode current no longer increases. In this state of saturation equation (5) is found to be exactly confirmed. The constant e appearing in the equation can be so accurately derived from the shot effect that this may be considered as one of the best methods of determining the charge on the electron.

Anode current fluctuations in amplifier valves

In the case of the customary amplifier valves (triodes and pentodes) the anode and the grid voltages are chosen so low that there can be no question of saturation. Otherwise indeed it would be impossible to regulate the anode current by variation of the control grid voltage. It is now found that the current fluctuations in the normal state of working of the amplifier valves are considerably smaller than would be expected according to equation (5). Qualitatively, this is easily understood. The electrons which are situated between cathode and control grid while the valve is in action (space charge) exert a repulsive effect on all electrons which are on the point of leaving the cathode. Because of this, part of the electrons emitted are driven back to the cathode, so that the anode current is much smaller than the emission of the cathode. When the cathode emission now fluctuates so that at a certain moment a larger number of electrons are emitted than the average, the space charge also becomes stronger and sends a larger percentage of the emitted electrons back to the cathode. The space charge thus smooths the current fluctuations. Experimental data on this smoothing in amplifier valves in practical use have already been given in this periodical [5]). The smoothing is expressed by a "noise factor" F, which indicates the ratio between the current fluctuations actually occurring and the current fluctuations which would be expected according to the shot effect. Expressed in a formula

$$F^2 = \frac{\overline{i_k^2}}{2\, e\, I_k\, \Delta f} \quad \dots \dots \quad (6)$$

The value of F^2 determined experimentally according to this formula is always less than unity in the case of ordinary amplifier valves, and in many cases it amounts to only a few per cent.

We shall now go somewhat more deeply into the theoretical basis of the smoothing of the current fluctuations by the charge.

Theory of the suppression of the anode current fluctuations by the space charge

A simple consideration, which makes it understandable that in the presence of a space charge the current fluctuations disappear in the first approximation, is possible on the basis of the theory of the space charge given by Langmuir [6]). Langmuir calculated for a diode with a plane cathode and a plane anode, the anode current per unit surface which may flow with a given anode voltage, when it is not the emission of the cathode but the repulsive action of the space charge which limits the current. The result is the well known equation:

$$I_k = \frac{1}{18\,\pi}\,\frac{m}{e}\,\frac{\left(\frac{2e}{m}\,V_a\right)^{3/2}}{d^2}, \quad \dots \dots \quad (7)$$

where e is again the charge and m the mass of the electron. V_a is the anode voltage and d is the distance between cathode plate and anode plate.

The important feature of this formula is that the saturation current I_s of the cathode does not occur in the result. The result is of course only valid when I_s is much greater than I_k.

Since the emission does not occur in the result it is also immediately clear that accidental fluctuations of the emission are unable to affect the magnitude of the current passing. If Langmuir's formula is correct, therefore, a complete suppression of the current fluctuations could be expected. As already mentioned the experimentally found suppression is indeed very strong. A certain residual effect remains, however, and indicates that the conception of the limitation of the current by the space charge as given here is too general and does not exactly correspond to the actual facts.

By means of a simple improvement in the calculations it is possible to explain the residual effect and to calculate it. In the derivation of equation (7) an error of neglect has been made by assuming that the electrons leave the cathode without initial velocity. Because of the fact that the electrons actually have a certain initial velocity, the space charge in the neighbourhood of the cathode is

[5]) See the second article mentioned in footnote [1]).

[6]) I. Langmuir, Phys. Rev. 2, 450, 1913; 21, 419, 1923.

slightly lower, so that the limit of the current prescribed by the space charge will be slightly greater.

The way in which the space charge limits the current is shown graphically in *fig. 3a*. Between cathode and anode a potential minimum occurs due

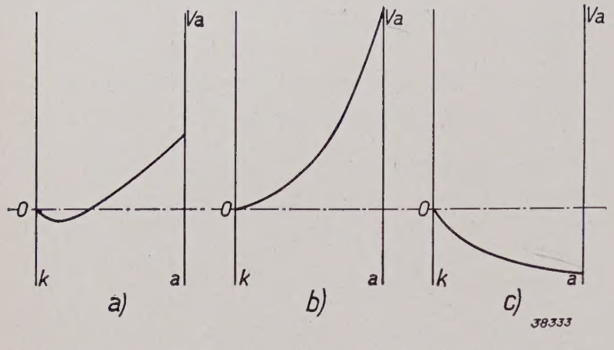

Fig. 3. Variation of the potential between cathode k and anode a of a diode, a) for the ordinary state of operation. b) for very high anode voltages, c) for very low (negative) anode voltages. In the ordinary state of operation the potential between cathode and anode exhibits a minimum.

to the space charge. Between the cathode and this minimum there is a retarding field for the electrons, which can only be overcome by those electrons whose initial velocity lies above a certain value. The other electrons return to the cathode. Furthermore, it may be seen from the figure that this minimum disappears for very high anode voltages (3*b*), as well as for sufficiently negative anode voltages (3*c*). In these limiting cases the regulating action of the space charge actually disappears, and the unimpaired shot effect is again obtained.

The correction of the initial velocity is easily added to Langmuir's calculation. One then finds for the first approximation of the current:

$$I_k = \underset{(u=0)}{I_k} \left(1 + \frac{3\bar{u}}{\sqrt{\frac{2e}{m}V_a}}\right), \quad . \quad . \quad . \quad (8)$$

where $\bar{u}$ is the average initial velocity of the electrons in a direction perpendicular to the cathode.

This velocity is determined by thermal factors; it corresponds approximately to the velocity which the electrons would possess in a gas at the temperature T_k of the cathode. It is known that the average kinetic energy of an atom or electron in a gas is equal to $^1/_2 kT_k$ per degree of freedom, so that $\overline{u^2}$ is of the order of magnitude of kT_k/m.

Like all thermal phenomena, the initial velocity exhibits certain fluctuations, and this is why the anode current is not absolutely constant. If we consider a certain time interval τ, the average velocity $\bar{u}_\tau$ of the electrons emitted in that time interval will in general differ from $\bar{u}$. If we set $\bar{u}_\tau - \bar{u} = \delta u$, then according to equation (8) the current fluctuations are:

$$\overline{i_k^2} = I_k^2 \frac{9\,\overline{\delta u^2}}{(2e/m)\,V_a} \quad . \quad . \quad . \quad . \quad . \quad . \quad (9)$$

The fluctuations of the initial velocity of the electrons are of the same order of magnitude as the initial velocity itself. If in each time interval τ one electron should be emitted, $\overline{\delta u^2}$, like $\overline{u^2}$, would amount to about kT_k/m. If n_τ is the average number of electrons which is emitted per time interval τ, the fluctuation of the *average velocity of these* n_τ *electrons* is smaller by a factor n_τ, so that one may write:

$$\overline{\delta u^2} = a\,\frac{kT_k/m}{n_\tau} = a\,\frac{kT_k/m}{I_k\,\tau/e}, \quad . \quad . \quad . \quad (10)$$

where a is a numerical factor of the order of magnitude of unity. An exact calculation gives: $a = 2(1-\pi/4) = 0.429$.

The square of the velocity fluctuations of the electrons is thus proportional to the cathode temperature T_k, and inversely proportional to the specific time interval τ. This is exactly the same relation as would be found for the thermal current fluctuations in a resistance if they were considered, not for a given frequency band, but for a given time interval. Conversely, in the formula for the velocity fluctuations we may also pass over from the time interval τ to a frequency interval Δf. A simple harmonic analysis shows that it is only necessary to substitute $2\Delta f$ for $1/\tau$ [7]). For the velocity fluctuations of the emitted electrons one then finds

$$\overline{\delta u^2} = 2\,a\,\frac{kT_k/m}{I_k/e}\,\Delta f$$

and by introducing this result in (9) one obtains the anode current fluctuations:

$$\overline{i_k^2} = I_k^2 \frac{9}{(2\,e/m)\,V_a}\,2\,a\,\frac{kT_k/m}{I_k/e}\,\Delta f =$$

$$= 9\,a\,kT_k\,\frac{I_k}{V_a}\Delta f. \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (11)$$

The most striking part of this result is that the properties of the electron (charge and mass) have disappeared entirely from the final formula. The current fluctuations of a diode whose anode current is limited by space charge are apparently a thermal phenomenon. The relation found for them very

[7]) The same transition from time intervals τ to frequency intervals Δf was carried out for the shot effect earlier in this periodical, and the following formulae were obtained:

$$\overline{i_k^2} = \frac{e\,I_k}{\tau} \quad \text{et} \quad \overline{i_k^2} = 2\,e\,I_k\,\Delta f$$

See the first article cited in note [1]).

much resembles the one for the thermal current fluctuations of a resistance (equation 1), since the quotient V_a/I_k may be considered as the internal resistance of the diode. For practical applications it is advantageous to introduce the slope $S = \mathrm{d}I_k/\mathrm{d}V_a$ of the diode instead of the quotient V_a/V_k. Since the anode current varies with the 3/2 power of the anode voltage

$$S = \frac{3}{2}\frac{I_k}{V_a} \text{ and therefore}$$

$$\overline{i_k^2} = 0{,}644 \cdot 4\, kT_k S\, \Delta f \quad . \quad . \quad . \quad . \quad (12)$$

Finally we may again express the fluctuations by the noise factor defined in equation (6). By combining (6) and (11) we then find:

$$F^2 = \frac{9\, a\, kT_k}{2\, e\, V_a} \, .$$

F is thus proportional to $1/\sqrt{V_a}$.

The result is only valid for anode voltages for which the approximations introduced during the calculations are permissible. This is the case only when V_a is not too low, but on the other hand far enough below the saturation value V_s. If the anode voltage lies outside this interval the calculation becomes much more difficult, and can only be carried out numerically [8]).

Fig. 4 gives a graphical idea of the result for a special case. The region in which our approximations are valid is indicated by dotted lines. Outside this region the noise factor increases suddenly; for negative anode voltages and anode voltages greater than V_s the noise factor is equal to unity.

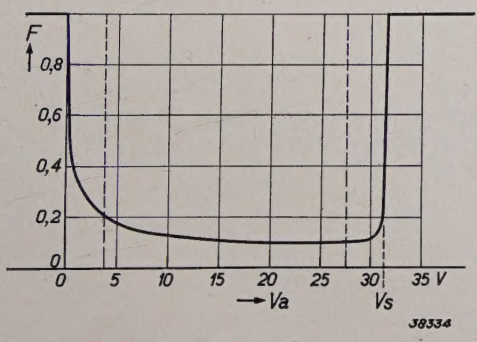

Fig. 4. Noise factor F as a function of the anode voltage. In the sloping part the variation is calculated according to equation (12), while in the neighbourhood of $V_a = 0$ and $V_a = V_s$ numerical calculations have been applied.

The results are well confirmed experimentally. In carrying out measurements on a diode larger values are usually found, it is true, than given by the theory, since electrons are reflected by the anode and disturb the field distribution and thereby cause additional fluctuations. If, however, a triode is used, these difficulties do not occur.

If equations (11) and (12) are applied to a triode the anode voltage must be replaced by the "effective control voltage" V_{eff} in the grid plane. This control voltage may be calculated from the anode current with the help of Langmuir's formula. The quantity S in equation (12) is now defined by $\mathrm{d}I_k/\mathrm{d}V_{\mathrm{eff}}$, *i.e.* the change in the anode current with the control voltage. If the anode voltage remains constant, the change in the control voltage is smaller than the change in potential of the grid wires themselves, since the anode potential also acts in the grid plane. The value of S to be used in equation (12) is therefore greater than the measured slope of the triode, the difference may amount to a factor 2, for instance.

Other fluctuation phenomena in triodes

The spontaneous fluctuations of the current which leaves the cathode are important not only because of their direct effect on the anode circuit, but also because of the fact that the alternating current upon passing the negative control grid wires induces an alternating electrical charge on the control grid. Since in general there is a very high impedance between control grid and cathode, the control grid current which corresponds to the fluctuations of the control grid charge may lead to considerable voltage variations. These variations in turn again cause fluctuations of the anode current. These fluctuations are found in the case of short waves to be able to furnish a very considerable contribution to the noise of receiving sets, and therefore require our attention [9]).

The negative charge Q which is present between the electrodes of an amplifier valve is given by the product of the current I_k and the transit time τ_{ka} of the electrons between cathode and anode. This charge leads to an equally large induced charge which is distributed over the different electrodes; a certain part of it q is therefore also present on the control grid. If the current I_k exhibits certain fluctuations i_k, the space charge $I_k\tau_{ka}$ fluctuates by an amount $i_k\tau_{ka}$, and we may therefore write for the fluctuations of the charge on the control grid:

$$\delta q = a\, i_k\, \tau_{ka}\, ,$$

where a is a factor smaller than unity.

[8]) E. Spenke, Wiss. Veröff. Siemens Werke **16**, 19, 1937.

[9]) C. J. Bakker, Physica **8**, 23, 1841.

In the case of amplifier valves of ordinary construction it is found that the induced charge on the control grid comes mainly from the electrons which are situated between cathode and control grid. If these electrons alone are taken into account, by substituting, instead of the whole transit time τ_{ka}, the transit time to the control grid τ_{ka}, one then calculates *as a first approximation* for a the value $^1/_3$, and one obtains the relation:

$$\delta q = \frac{1}{3}\, \dot{i}_k\, \tau_{kg} \quad . \; . \; . \; . \; . \; . \quad (13)$$

We shall now imagine the current fluctuations to be resolved into components of different frequencies, as in the foregoing, and shall consider only those components for which the angular frequency lies in the neighbourhood of a certain value ω. The variation δq then leads to a control grid current of the following magnitude:

$$\dot{i}_g = \frac{1}{3}\, j\omega\tau_{kg}\, \dot{i}_k \quad . \; . \; . \; . \; . \quad (14)$$

and for the mean square of the current fluctuations one thus obtains:

$$\overline{i_g^2} = \overline{i_k^2}\, \frac{1}{9}\, \omega^2\, \tau_{kg}^2.$$

If one now substitutes for $\overline{i_k^2}$ in this equation the value according to equation (12), one obtains:

$$\overline{i_g^2} = \frac{1}{9}\, 0{,}644 \cdot 4\, kT_k\, S\Delta f \cdot \omega^2 \tau_{kg}^2 \; . \; . \quad (15)$$

Finally the result may be simplified by making use of the fact that the transit time of the electrons leads not only to the grid current just calculated, but also to a damping in the grid circuit. The explanantion and calculation of this "electron input damping" have already been given in this periodical [10]). The result found is:

$$\frac{1}{R_e} = \frac{1}{20}\, S\omega^2 \tau_{kg}^2.$$

The value of R_e thus obtained is called the electron input resistance of the valve. By substituting the electron input resistance in equation (15) one finds:

$$\overline{i_g^2} = \frac{20}{9} \cdot 0{,}644 \cdot \frac{4\, kT_k}{R_e}\, \Delta f = 1{,}43\, \frac{4\, kT_k}{R_e}\, \Delta f \; . \quad (16)$$

Thus again a formula which closely resembles the one for the thermal current fluctuations of a resistance. We are, however, of the opinion that the numerical difference (in this case the factor 1.43) is essential, and shows clearly that we are here not concerned with a phenonenon which could have been derived by a thermodynamic method.

The grid current fluctuations are determined by the cathode temperature T_k which is about 4 times as high as room temperature. Together with the factor 1.43, therefore, one finds that the electron input resistance causes current fluctuations at least 5 times as large as a resistance of the same size which is connected externally between cathode and control grid. At very high frequencies the electron input damping $1/R_e$ is often greater than the damping $1/R_{LC}$ of the oscillator circuit in parallel with it. The spontaneous current fluctuations in the grid circuit may then form the most important source of disturbance of the receiving set.

The result, equation (16), was tested by means of experiments with an amplifier valve of the type EF 50. As may be seen in *fig. 5*, the grid current fluctuations $\overline{i_g^2}$ are actually proportional to the electron input damping $1/R_e$; they both vary proportionally with ω^2, as the theory demands. The absolute magnitude of the current fluctuations measured also agrees well with the calculated values (see *table I*).

Table 1

Calculated and measured values of the grid current fluctuations $\overline{i_g^2}$ for the three working states indicated in fig. 5 of the pentode EF 50.

I_a (mA)	$\overline{i_g^2}$ (Amp.^{2})	
	measured	calculated
1.20	$1.9 \cdot 10^{-41}\ \omega^2\, \Delta f$	$1.6 \cdot 10^{-41}\ \omega^2\, \Delta f$
4.15	$4.8 \cdot 10^{-41}\ \omega^2\, \Delta f$	$4.5 \cdot 10^{-41}\ \omega^2\, \Delta f$
9.60	$9.3 \cdot 10^{-41}\ \omega^2\, \Delta f$	$7.4 \cdot 10^{-41}\ \omega^2\, \Delta f$

ω angular frequency, f frequency, both in sec^{-1}.

Current fluctuations in valves with more than one positive electrode

a) *Distribution fluctuations*

In modern receiving sets the first amplifier valve is almost always a screen grid valve, *i.e.* a valve

[10]) The electron input may be explained as follows. If a grid A.C. voltage of not too high a frequency is applied between the control grid and cathode of an amplifier valve, an alternating charge in the same phase is present on the grid, which is manifested in a capacitative grid current 90° in phase ahead of the grid voltage. At very high frequencies, however, it must be taken into account that the charge on the control grid consists partially of the induced charge of the electrons between control grid and cathode. This induced charge follows the voltage variations with a certain time lag which is determined by the transit time of the electrons. The result is a phase shift of the induced current, which thus no longer remains purely capacitative, but takes on a real component in phase with the applied voltage. The calculation of this component, the electron input damping, is outlined in Philips techn. Rev. 1, 176, 1936.

which contains a screen grid in addition to the control grid. The screen grid, which is placed between control grid and anode, is at a constant positive potential and in this way serves to prevent (undesired) changes in the potential field in the neighbourhood of the control grid caused by changes in the anode voltage.

The electrons which pass from the cathode to the anode must therefore pass the screen grid as well as the control grid. Since the former is also at a positive potential, part of the electrons are captured by the grid wires. As a result of this a shot effect again occurs, *i.e.* even when the number of electrons which moves toward the screen grid per time interval τ is absolutely constant, the number of elec-

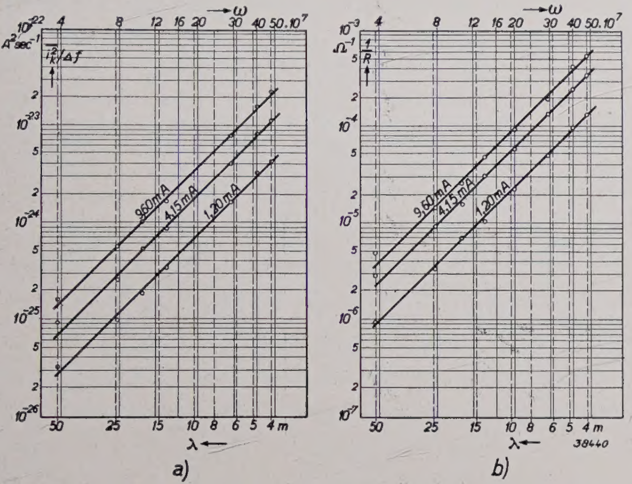

Fig. 5. *a*) Grid current fluctuations $\overline{i_g^2}$,*) *b*) input damping $1/R$ of a pentode EF 50 as a function of the frequency (the figure does not give the electron input damping, but the total input damping). The curves were recorded for three different adjustments of the value with anode currents of 1.20 mA, 4.15 mA and 9.60 mA, respectively. It may be seen that both $\overline{i_g^2}$ and $1/R$ are proportional to ω^2.

trons which impinges on the wires of the grid instead of passing between them will still exhibit certain fluctuations. This causes current fluctuations in the screen grid circuit, and at the same time — with an opposite sign — in the anode circuit. For the size of the fluctuations, the so-called current distribution fluctuations, one may expect equation (5) of the shot effect to be approximately valid:

$$\overline{i_{g2}^2} = 2\, e\, I_{g2}\, \Delta f.$$

This result is only valid when the cathode current exhibits no fluctuations, and when, moreover, the screen grid current I_{g2} is small compared with the total cathode current $I_k = I_{g2} + I_a$. If the cathode current fluctuations given by a noise factor F are included in the calculation, one finds for the total fluctuations in the anode circuit the formula [11]):

$$\overline{i_a^2} = 2\, e\, \frac{I_a}{I_k}\, (I_{g2} + F^2 I_a)\, \Delta f, \quad . \quad . \quad (17)$$

where the first term in parenthesis refers to the distribution fluctuations and the last term to the cathode current fluctuations.

In valves of ordinary construction the screen grid current may amount for instance to 20 per cent of the anode current. The square of the current distribution fluctuation $\overline{i_{g2}^2}$ then also amounts to about 20 per cent of the current fluctuations which the anode current would exhibit in the case of unimpaired shot effect. Now we have seen that the cathode current fluctuations can be weakened by the space charge to less than 10 per cent of their original value. The distribution fluctuations then form the chief source of noise disturbance in the anode circuit.

b) Fluctuations due to secondary emission

In order to increase the slope $\mathrm{d}I_a/\mathrm{d}V_g$, of radio valves, more and more use is being made of the phenomenon of secondary emission. *Fig. 6* shows diagrammatically the arrangement of the electrodes in such a valve. The current I_{prim} strikes the auxiliary electrode h which is covered with a secondary emitting substance. The current of secondary electrons is received by the anode a. This current has a value $I_{\text{sec}} = \delta I_{\text{prim}}$ when δ is the average number

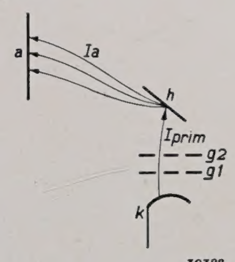

Fig. 6. Electrode system of an amplifier valve with secondary emission.

of secondary electrons liberated per primary electron from the auxiliary electrode.

Like the direct current, the slope is increased by a factor δ. Since the factor δ may amount for instance to 4.5, this is no inconsiderable improvement.

The fluctuations of the anode current may be written in the form:

$$\overline{i_a^2} = \overline{i_{\text{prim}}^2}\, \delta^2 + \overline{i_{\text{sec}}^2}, \quad . \quad . \quad . \quad . \quad (18)$$

The first term of this equation indicates that the fluctuations of the primary current are increased by a factor δ. The second term, $\overline{i_{\text{sec}}^2}$, is due to the fact

*) In the figure by error $\overline{i_k^2}$ has been used.

[11]) C. J. Bakker, Physica **5**, 581, 1938.

that a primary electron which arrives on the auxiliary electrode does not always free the same number of secondary electrons, but that in this phenomenon also certain fluctuations occur.

The number of secondary electrons per primary electron may be described by a statistical distribution around an average value. Let $\beta_0, \beta_1 \dots \beta_m \dots$ be the chance that 0,1, ... m ... secondary electrons are freed by a primary electron ($\sum_m \beta_m = 1$), then $\delta = \sum_m \beta_m m$.

It is obvious that a primary electron which frees m secondary electrons will lead to a current fluctuation proportional to $m-\delta$. The absolute value of this current fluctuation i_m is given by an expression which closely resembles the familiar formula for the shot effect, namely:

$$\overline{i_m^2} = 2\, e\, I_{\text{prim}}\, \beta_m\, (m-\delta)^2\, \Delta f.$$

The total contribution to the fluctuation $\overline{i_{\text{sec}}^2}$ is obtained by a summation of this result over all values of m. If, further, we write for the fluctuation of the primary current,

$$\overline{i_{\text{prim}}^2} = 2\, e\, I_{\text{prim}}\, F_{\text{prim}}^2\, \Delta f,$$

we obtain

$$\overline{i_a^2} = 2\, e\, I_{\text{prim}}\, [F_{\text{prim}}^2\, \delta^2 + \sum_m \beta_m\, (m-\delta)^2]\, \Delta f \quad . \quad (19)$$

In order to apply this formula, for a given material and a given velocity of the primary electrons, one would have to know the probabilities β_m for each value of m. An experimental determination of these probabilities is impossible at present. From measurements on actual valves it is found that the second term of equation (19) may be of the same order of magnitude as the first term, so that the extra current fluctuations caused by the secondary emission furnish no unimportant contribution to the total fluctuations of the anode current.

The last relation is very much simplified if the primary current exhibits the true shot effect [12]), so that $F_{\text{prim}} = 1$. Then according to equation (19):

$$\overline{i_a^2} = 2\, e\, I_{\text{prim}}\, (\sum_m \beta_m\, m^2)\, \Delta f, \quad . \quad . \quad . \quad (20)$$

an equation which can also be directly understood by considering that every term of the sum in parenthesis represents an anode current $I_{\text{prim}} \cdot \beta_m \cdot m$, which consists in each case of portions of m electrons, so that it will furnish a contribution to the shot effect, corresponding to an elementary charge $m.e$.

[12]) M. Ziegler, Physica 3, 1, 1936.

NEW PRINCIPLES OF CONSTRUCTION FOR THE ELECTRO-ACOUSTIC INSTALLATION OF STUDIOS

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by F. DE FREMERY and J. W. G. WENKE.

The studio building which was put into use as an annex of the old A.V.R.O.-studio building in 1940 possesses an installation which differs from the customary one in various important features. The contact between the performing artists and the technical personnel has been made closer than was previously the case. The amplifiers occurring in the installation are not fed from a central battery, but each one is fed from its own built-in supply apparatus. By the standardization of levels and impedances of the connection lines, the number of different types of amplifiers could be reduced to three. Inputs and outputs of the amplifiers can be connected by means of cross connection panels to the different microphones, regulators, lines, etc., whereby with a relatively small number of amplifiers all the possibilities of connection occurring during use can be realized.

A description has already been given in this periodical [1]) of the large studio building of the A.V.R.O. which was inaugurated in 1936. After only a few years, however, the building was already found inadequate for the ever increasing activities connected with broadcasting, so that a start was made on the construction of a new annex studio building. This building which contains two concert studios, and which was finished last year, deviates considerably in arrangement from the old building. In particular a number of important new principles have been employed in the electro-acoustic system, which, like that in the older building, was installed by N.S.F.-Philips. The most important points will be discussed in the following.

The connection of the control tables

There is a certain two-sidedness in the activities of a broadcasting studio: at least in the main programmes the object is the achievement of artistic performances, while on the other hand purely technical questions are always in the foreground. If, for example, we consider the broadcasting of a concert where the sound of different parts of the orchestra and possible soloists is taken up by different microphones, the different microphone contributions must then be regulated separately and mixed in the correct proportions, and at the same time the dynamics of the sound must be adapted to the difficulties of the broadcast, *i.e.* the amplification of the whole must continually be regulated during the performance in order to bring out the pianissimos sufficiently strongly above the natural interference level (noise) and in order not to exceed the maximum permissible depth of modulation in the fortissimos [2]). While the artistic control is in the hands of the conductor, there is a very real danger that his intentions will be nullified by these technical processes, for instance, that the balance of orchestra and soloists will be disturbed or dynamic finesses regulated out of existence.

In order to avoid this, in addition of course to a good training of the mixing and regulating technicians, a visual contact between the latter and the conductor is also desired. In the old A.V.R.O. building the necessity was met by glass cabins in the studios, in which the person is seated who mixes the contributions of the different microphones in the studio in question. The regulation of the whole broadcast then takes place at a control table, which is placed in a separate room (*fig. 1*), and on which, in addition to the control instruments (modulation meter and monitor loud speaker), the regulating

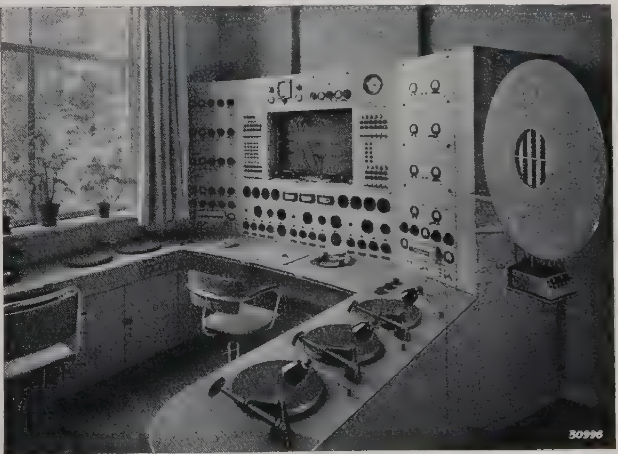

Fig. 1. A central control table in the old A.V.R.O. studio building. The person at this table had no visual contact with the studio. In addition to regulators and control instruments various auxiliary apparatus are installed on the table (gramophones, amplifiers, etc.).

1) The equipment of broadcasting studios, Philips techn. Rev. **4**, 136, 1939.

2) See for example R. Vermeulen. The relationship between fortissimo and pianissimo, Philips techn. Rev. **2**, 266, 1937.

elements and the necessary signalling systems, there is also a series of auxiliary apparatus, such as gramophone turntables, amplifiers, a switchboard for making the necessary connections between microphones, regulators, line to the transmitter, etc. This arrangement and the division of functions thereby involved between the collaborating technicians was chosen in order to permit as manysided use as possible of the control tables, for instance, for the broadcasting of radio plays in which different studios sometimes work together.

In the new studio building, which is intended only for the broadcasting of concerts, more emphasis could be laid on the collaboration between conductor and technicians. For each of the two studios (a large concert hall and a smaller studio for dance music) the mixing cabin and the control table are as it were combined, by bringing the control table itself into visual contact with the studio. At the same time the mixing and regulation and the whole direction of the broadcast are in the hands of a single person, who, by means of direct contact with conductor and orchestra, can follow the performance in all its details. The way in which the control tables for the two studios are arranged is shown in the floor plan of *fig. 2*, while *fig. 3* shows the control table which gives upon the large concert hall (*fig. 4*). In order to keep the view of the hall entirely free the table is low and only the operating elements are included, while the above-mentioned auxiliary apparatus of both tables are housed together in a separate room (the "technical service", see fig. 2). A more detailed description of the arrangement of the control tables is given in the text under fig. 3.

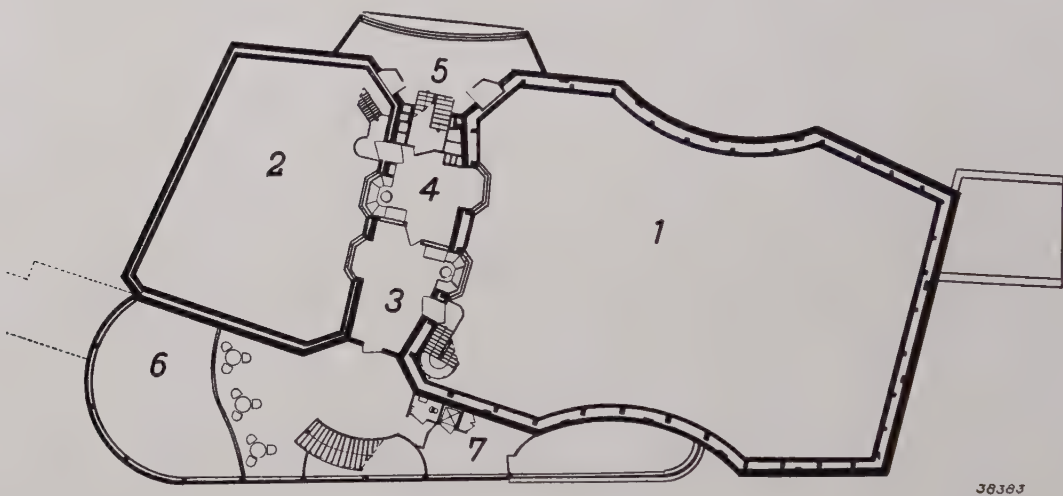

Fig. 2. Ground plan of the new A.V.R.O. studio building containing two studios: a large concert hall (*1*) and a smaller studio for dance music (*2*). Each studio has a control table. The two control tables are housed in cabins (*3*, *4*) between the two studios and have a free view of the studios *via* triple glass windows which furnish adequate acoustic insulation. *5* technical service room, *6* hall, *7* conductor's room. The unusual floor plan of the large concert hall, which recalls somewhat the sound box of a violin, has been chosen for acoustic reasons. As may be seen the two studios are constructed as entirely free-standing boxes inside the building proper, with their own walls and a separate foundation. This separation between studio and outside world, which was also applied in the old A.V.R.O. building, and which has been carried through as far as possible, insures a very good acoustic insulation. (The figure is borrowed from "De 8 en Opbouw" **11**, 173, 1940).

The supply of the installation

Roughly speaking, the electro-acoustic installation of a studio building is nothing but a combination of a large number of amplifiers which must supply all kinds of low-frequency voltages at different levels to regulators, lines and loud speakers. For the general arrangement of a studio installation the article cited may be referred to [1]). Without entering into details we shall here consider a few essential points, the first of which is the supply.

The feeding of the many amplifiers, both with heating currents and anode voltages, usually took place from a central accumulator battery; this was the case in the old A.V.R.O. building. The battery there set up was, however, not large enough to supply the installation of the new building as well, so that the problem of supply had to be considered anew.

Now battery feeding of itself has several objecjections. Apart from the high cost of installation and the complicated maintenance, it is particularly the mutual coupling of the amplifiers connected

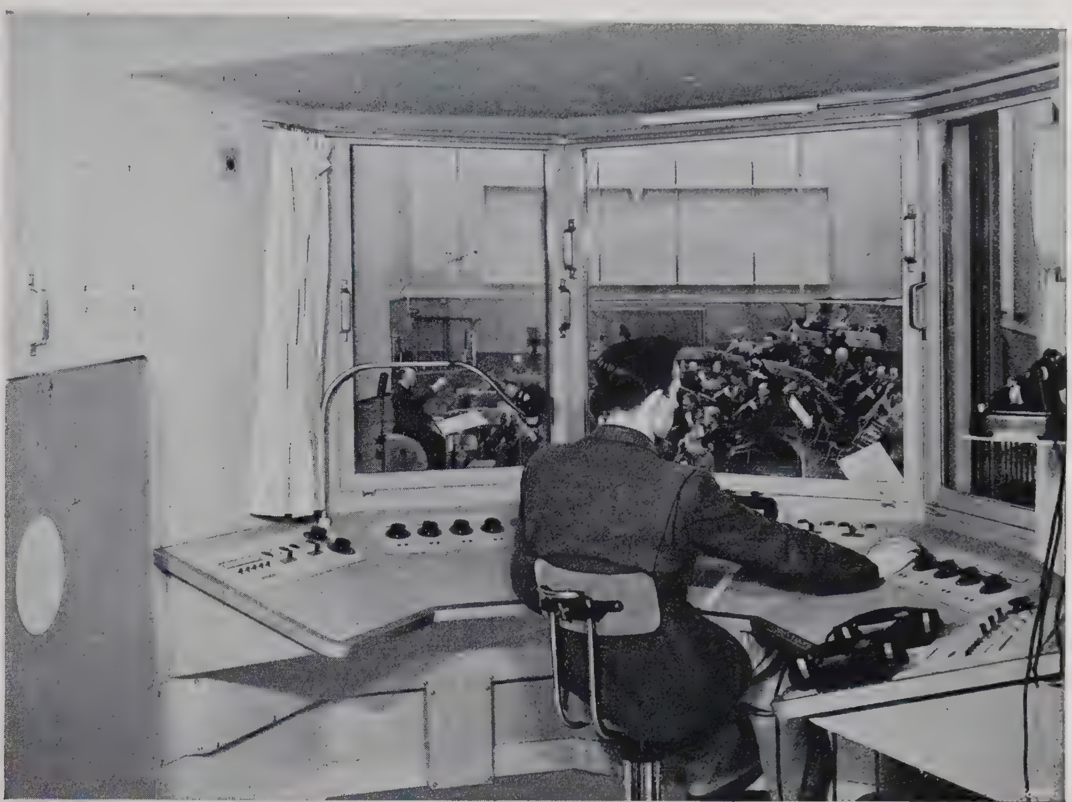

Fig. 3. Control table for the large concert hall. Due to the fact that all auxiliary arrangements not directly needed during the performance are housed elsewhere (in the Technical Service room, *5* in fig. 2), the table could be so arranged that an uninterrupted view of the hall is possible. On the five sloping panels are to be found from left to right: 1) switches and signal lamps for connections to the old building, a switch and a regulator for connecting the monitor loud speaker (upper right) to different studios, a microphone for giving directions to the studios, a set of knobs by means of which indications on a light screen can still be given to the conductor after the beginning of the broadcast. 2) A group of four microphone regulators for mixing four microphone contributions and a main regulator for regulating the whole. 3) Switches and signal lamps for signalling in the two studios and the corresponding announcer's cabins, and the indicating instrument of the modulation meter, with which the peak voltages in the programme are controlled. 4) Four more microphone regulators and a main regulator. 5) Telephone apparatus, switches for programme contributions from outside the studio, for instance for mixing sound effects (which can here be furnished from a separate studio of the old building), as well as keys and knobs for signalling to the second control table.

to the central battery which forms an objection. It is difficult to make the internal resistance of the battery so small that the coupling is sufficiently restricted, and recourse must therefore be taken to the introduction of decoupling circuits. Considering the very low frequencies (30 c/s) which must be dealt which in the amplifiers, the decoupling circuits must contain very large self-inductances [3]), which in turn involves considerable current surges upon switching on and off of an amplifier. These current surges lead to click disturbances in all the amplifiers connected.

In addition to these "natural" disadvantages of battery supply, in our case there was also the fact that it was a very uneconomical solution for the relatively limited installation of the new building to set up a second separate central battery. This was the reason why a different solution of the problem was applied, namely the supply of each amplifier separately by a built-in supply apparatus.

The fact that this solution, in which practically all coupling between the amplifiers is eliminated, has been applied until now only very seldom, is due mainly to the fear of the hum caused by the 50 c/s of the A.C. mains. This is indeed a difficulty which must not be underestimated, especially in a studio where all the amplifiers must be able to pass the frequency of 50 c/s in full strength, and where partially very weak signals occur which are therefore sensitive to disturbances. The problem could, however, be solved satisfactorily in the following way.

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[3]) In telephone networks where the same problem occurs — cross talk between the microphone connected in parallel on a battery — the problem is much simpler, since practically only frequencies above 300 c/s. are encountered.

Fig. 4. View of the large concert hall. Two hanging microphones on rails may be seen. Next to the position of the conductor is a movable signal box with a light screen on which the director of the broadcast, seated at the control table, can give several commonly occurring indications to the conductor. Right, sound effect loud speaker for the mixing of background noises. The numerous panels of the walls can be set at different angles in order to influence the acoustics of the hall (sound distribution, reverberation).

In order to limit the hum to the required degree the windings of input and output transformers of each amplifier must be especially protected against induction by spread lines of force from the supply transformer. Various means of doing this have been applied. In the first place the supply transformer works with a lower iron induction, *i.e.* not so close to saturation as is usually the case. This gives a higher permeability of the iron and therefore less spreading. Moreover, the supply transformer is placed in a heavy cast-iron pot which tends to absorb the spread lines of force and thus diminishes still more the field observable toward the outside. Furthermore the input and output transformer are placed in the amplifier panel as far as possible outside the sphere of the field of the supply transformer. This may clearly be seen in *fig. 5*: the supply transformer is at the upper left hand, the input transformer at the extreme right, the input transformer about in the middle at the bottom (because it is somewhat less sensitive to hum than the input transformer due to the greater signal intensity at that point). The input and output transformers are also housed in pots for shielding. Since it is

here a question of shielding fields which are themselves quite weak, an alloy has been used for these pots which has a high permeability especially at low induction. Moreover, for the very sensitive the earth currents to follow all kinds of complex routes, but on the contrary they are insulated from the panel and connected to a common earth point of the amplifier by the straightest possible connections.

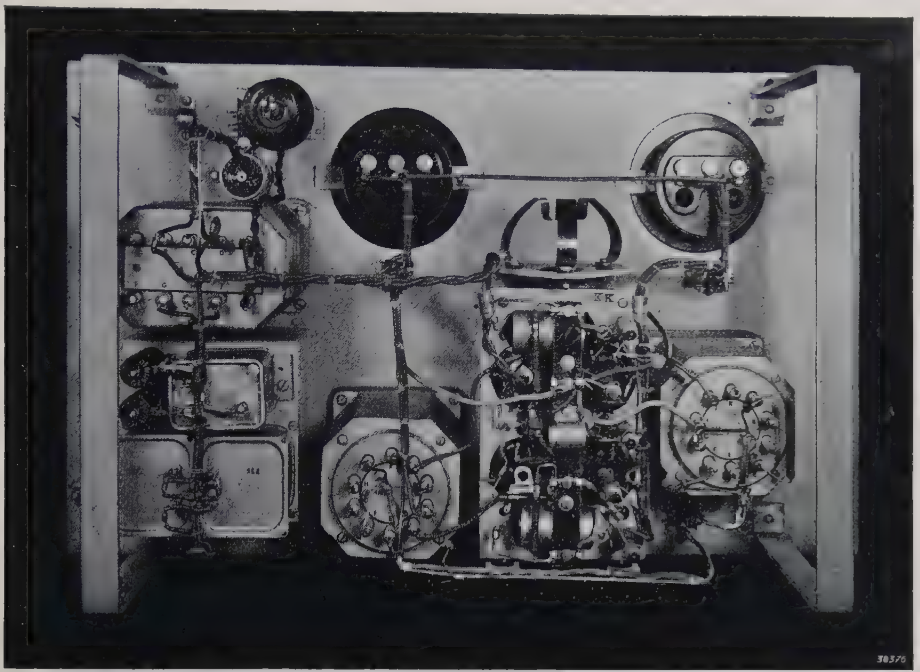


Fig. 5. Arrangement of the components in the panel of an amplifier. The supply transformer on the one hand (left) and the input and output transformer on the other hand (right and below) are kept as far as possible away from each other. Above the supply transformer may be seen a "Starto" valve (automatic starting resistance [4]), which limits the initial surge and thereby helps to prevent clicking in neighbouring amplifiers.

input transformer an additional shielding inside the pot of a special kind of sheet metal is used, while its winding is "astatic", see *fig. 6*. By this means the sensitivity for external fields is made very small. An external field gives in general lines of force of the same direction in the two arms of the transformer core, in the two halves of the secondary winding therefore electromotive forces in the same direction are induced, and since the two halves are connected in opposition, the resulting voltage is zero. Nevertheless the transformer works quite normally for a primary current, because the latter causes a magnetic field whose lines of force run in opposite directions in the two arms of the core, so that the induced electromotive forces in the two halves of the secondary winding are added together.

Special attention is also paid to the earth connection in order to decrease the hum. The transformer pots, which are to a certain degree coupled capacitatively with the windings, are not connected electrically to the panel, since this would enable

The cathodes of the amplifier valves are indirectly heated.

By these and other measures the hum could be sufficiently suppressed. A measurement of the hum and noise voltage [5]) in the case of the amplifier

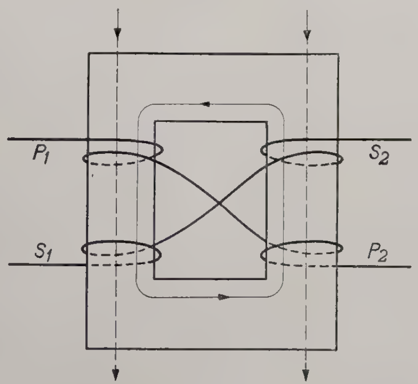


Fig. 6. Astatically wound transformer. P_1, P_2 halves of the primary, S_1, S_2 halves of the secondary winding. The primary current causes in the two arms of the core opposite lines of force, an external field, on the other hand, causes lines of force in the same direction (broken lines).

[4]) See P. C. van der Willigen, Philips techn. Rev. 1, 205, 1936.

[5]) The measurements were carried out with a so-called psophometer which takes into account the dependence on frequency of the sensitivity of the ear.

type *I* gave a value of 0.5 mV at the output, *i.e.* about 75 dB below the normal line level.

Standardization of the amplifiers

In older studio installations, including that of the old A.V.R.O. building, amplifiers of a different type were used for each function; thus there are microphone pre-amplifiers, gramophone amplifiers, programme amplifiers, coupling amplifiers, monitor amplifiers, etc. The differences between these types lie in the amplification produced, the input impedance, the output, etc.

This specialization and close adaptation of each amplifier to a certain purpose has disadvantages in manufacture as well as in use. In manufacture, because so many types of amplifiers must be made; in use, because so many amplifiers are needed, only a few of which are usually in action at one line, and for which the necessary reserves must be at hand. In our case, where each amplifier must also have a separate supply apparatus, these disadvantages were doubly felt.

Therefore a standardization of the amplifiers was undertaken for the new installation. All microphone connections, gramophone pick-ups, regulators are adapted to an input impedance of 200 ohms of the amplifier to be connected; by giving this an amplification sufficient for all purposes (60 dB), a single type of amplifier (*I*) proved sufficient. Only two other types of amplifier are then still needed for the listening to the broadcast at various spots with monitor loud speakers. The first is a coupling amplifier (*II*), which taps off the listening lines from the programme line (line to the transmitter) in such a way that commutations in the listening line have no reaction on the programme line, and which has a sufficiently high input impedance (5 000 ohms) not to affect the level of the programme line. The last is a power amplifier (*III*) which delivers the necessary power to supply a loud speaker (7 W). In fig. 10 at the end of this communication the way in which the three types of amplifiers are used may be seen.

All three types of amplifiers are two-stage push-pull amplifiers. By the push-pull connection all the connections can be kept symmetrical with respect to earth, which restricts the possible disturbances. Furthermore, due to the push-pull connection, the non-linear distortion is slight: the maximum distortion factor of the amplifiers of type *I* is 1 per cent above 100 c/s and 2 per cent below 100 c/s. In order to control the balance each amplifier stage contains a differential meter which indicates the difference current of the two valves.

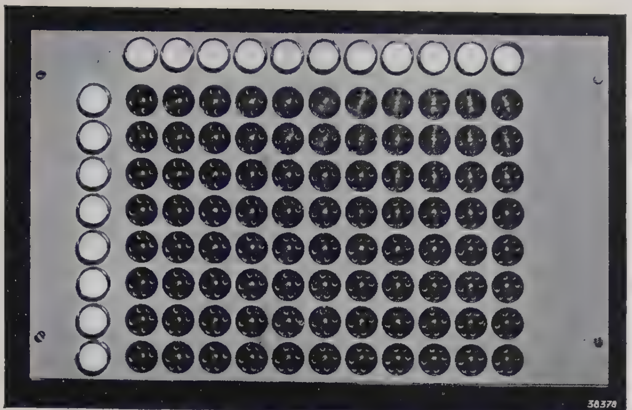
Cable installation and cross connection panels

For the manufacture the desired simplification is simply obtained with the standardization of the amplifiers. In order to be able to take advantage of this in use also, amplifiers *I*, which can be used in various branches, are not included rigidly in certain lines, but the inputs and outputs of all these amplifiers are brought together on switchboards where all the desired connections can be made. The number of amplifiers of type *I* could thereby be limited to the relatively small number of 14, whereby each amplifier may act as a reserve for every other amplifier in case of defect. In *fig.* 7 may be seen the racks in which these 14 amplifiers are placed.

An important consequence of this centralization of the amplifiers is that the microphones in the studios now possess no pre-amplifiers situated in the immediate neighbourhood, but are connected by a relatively long connection to the amplifiers. This made it necessary to pay special attention to the combination of interferences with the weak microphone voltages. In the first place in designing the studio building more care had to be taken than in previous buildings that the routes of the

Fig. 7. Rack in the technical service room in which the 14 amplifiers *I* are placed. On the panel of each amplifier may be seen the two differential meters for controlling the balance, and a small door giving access to the valves.

a



b

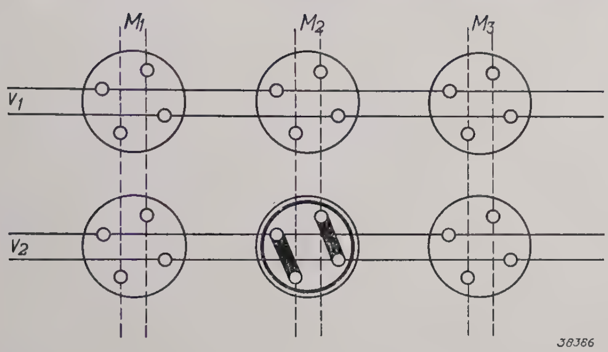


Fig. 8. *a*) Panel of cross connection units, with signal lamps at the ends of the horizontal and vertical rows.

b) Method of connecting the contact sockets of the cross connection units. M_1, M_2, M_3, etc., go to microphones and the like, V_1, V_2 etc. to amplifier inputs. A plug inserted for instance at the crossing point of V_2 and M_2 makes the connection there indicated.

"sensitive" microphone cables were kept away from those of the connections for the lights, the house telephones, the signalling, the electric clocks, etc. A plane at the height of the floor of the studios is therefore reserved exclusively for the microphone lines; the microphone connections are also at this height in the base boards of the studios. At places where the microphone lines must depart from this height, for instance for the connection with the points of suspension of the two hanging microphones in the large concert hall (fig. 4) a wide space is left free around the lines. All the connection lines are in continuous carefully earthed lead coverings for the sake of good shielding.

In the second place, the switching of the different microphone connections (14 in all) to the various amplifier inputs requires attention. The commonest method used in such cases is the switchboard, where each of the m outgoing connections is represented by a socket and each of the n incoming connections ends in a cord and a jack which can be plugged into any of the m sockets. This method could not be used here because of the very low level of the signals occurring at the amplifier inputs: the free hanging cords might here too easily lead to movements in the contacts and thus to crackle disturbances. A different system was therefore used, the cross connection panel. This can be pictured most simply as a set of m horizontal and n vertical rails crossing each other. Each horizontal rail is connected with an amplifier input, each vertical rail with a microphone line. By means of a jack a connection between two rails can be made at every crossing point.

Actually the two sets of rails are replaced by a panel of so-called "cross connection units", see *fig. 8*. Each connection unit contains two "horizontal" and two "vertical" contact sockets which are connected in the way shown in fig. 8*b*. The plugs which are inserted in these cross connection units (*fig. 9*) contain four very flexible contact pins of a special construction, which are connected in twos. In this way very good connections are obtained free of interference. In order to avoid mutual influences of microphones lines and amplifier inputs which are not connected with each other, there is a shielding in each cross connection unit between the "horizontal" and the "vertical" contact sockets. Furthermore the cross connection units also contain a number of auxiliary contacts (see fig. 9) which can be worked by two long auxiliary pins of the plugs and which may be used for signalling and locking. For instance, a series of signal lamps are operated by this means. These lamps are mounted around the cross connection panel (see fig. 8), when a plug is inserted at a crossing point the lamps at the ends of the horizontal and vertical rows involved are lighted to indicate the connection made and as a warning that these rows may no longer be used for other purposes.

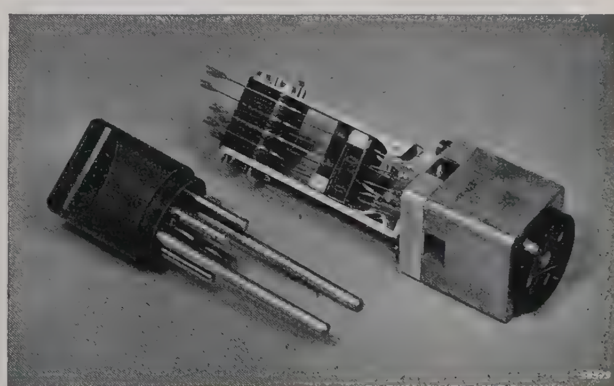


Fig. 9. Cross connection unit (right) and plug (left). On the body of the cross connection unit which contains two pairs of mutually shielded contact sockets may be seen the groups of springs of the auxiliary contacts which are operated by the two long auxiliary pins of the plug. The four short main pins of the plug are of a special very springy construction.

The outputs of the 14 amplifiers *I* also, with the lines to the regulators, the control instruments, the transmitter, etc. which must be connected to them are brought together on a cross connection panel. In fig. 7 it may be seen how the input and the output cross connection panels are placed to the left and right of the racks with the 14 amplifiers. The signalling system just described is so constructed that upon inserting a plug in the input cross connection panel the lamp at the corresponding horizontal row of the output cross connection panel also lights up faintly in order to facilitate the finding of the correct spot for plugging in on that panel, and at the same time to indicate again that the amplifier in question is in use.

In conclusion in *fig. 10* a view of the complete installation is given. On the cross connection panels as an example the connections are made for a broadcast in which three microphones in the studio collaborate (main regulator H_1), while an acoustic background (sound effect) is provided by the gramophone turntable which can be added to the programme *via* the main regulator H_2. At the same time this background can be sent to the studio itself *via* the regulator *R*. The communications of the announcer can also pass over the main regulator H_2. Several details are further explained in the text under fig. 10.

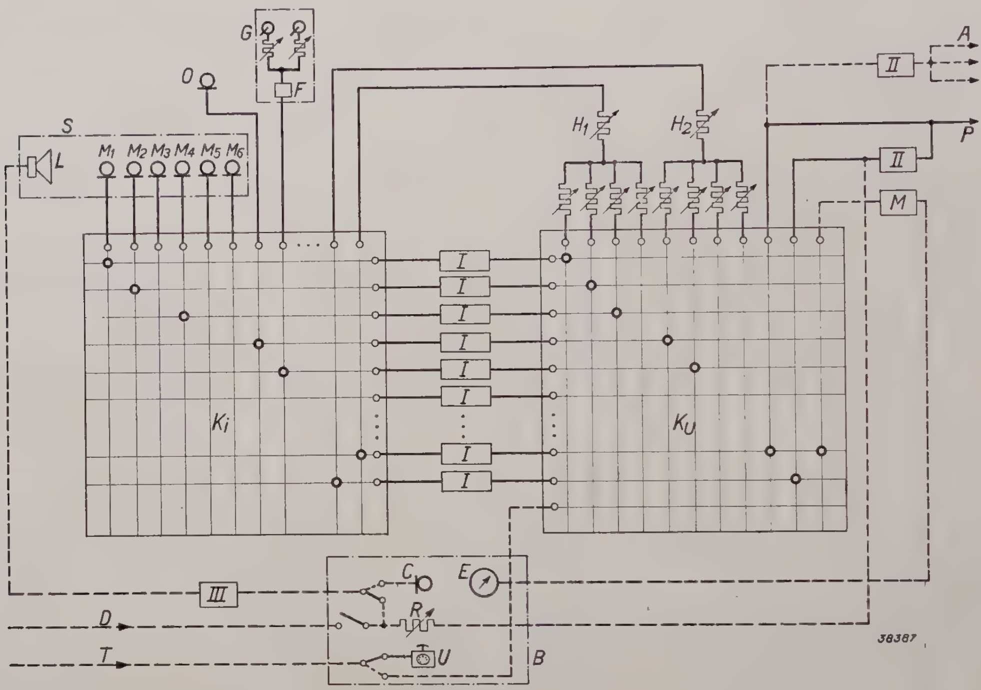


Fig. 10. View of the electro-acoustic installation of the new studio building. K_i input cross connection panel, K_u output cross connection panel, *S* studio with six microphones M_1-M_6 and sound effect loud speaker (at the same time loud speaker for giving directions); *O* microphone in the announcer's cabin, *G* gramophone turntable with correction filter *F*, H_1 and H_2 main regulators, *P* programme line, *A* monitor lines, *I* universal amplifier, *II* coupling amplifier, *III* power amplifier, *M* modulation meter with indicating instrument *R*; *B* cabin with microphone for giving directions *C*, regulator *R* for sound effect which can be supplied over line *D* from the old building, *U* telephone apparatus connected to the telephone line *T*, over which contributions to the programme may also be brought in.

THE CALCULATION OF LIGHTING INSTALLATIONS WITH LINEAR SOURCES OF LIGHT

by H. ZIJL.

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Several diagrams are derived which make it possible to determine in a simple way the illumination, and from it the luminous efficiency for installations equipped with linear light sources.

Introduction

In designing a lighting system with ordinary electric lamps and normal fittings one can make use of extensive factual material, of a theoretical as well as of an empirical nature, about the distribution of the illumination with the customary arrangement of the light sources. In the literature more or less detailed information can be found, which is usuallly arranged in the form of tables[1]) and in general is sufficiently accurate for practical use. The modern linear sources of light, however, such as the "Philora" TL lamps. are entirely different from the lamps generally used until now with respect not only to the distribution of their luminous flux but also to the way in which they are installed in the rooms to be illuminated, so that the experience which has been gained with earlier lighting systems cannot be directly applied to them. The number of systems already installed with such linear light sources is still too small to make it possible to collect adequate experimental data from them. It is therefore necessary for the present to calculate the distribution of the luminous flux for linear light sources theoretically, as was also formerly the case for ordinary systems with ordinary electric lamps. From this the so-called luminous efficiency of the system can easily be found. This is of course a more elaborate method than consulting a table, but it is unavoidable until such practical tables have been crystallized out of many calculations confirmed by illumination measurements. Moreover, it offers the opportunity of attaining a clearer insight into the problems in question than is usually obtained by the mechanical calculation with tables.

Illumination systems with linear light sources usually consist of tubes which are mounted horizontally close under the ceiling with or without reflectors. Such a system must then provide the greatest possible illumination on the tables present in the room, which generally have their working surface about 1 m above the floor. The luminous flux from the lamps may be divided into three parts (*fig. 1*) which are incident:

a) directly on the working surface,
b) directly on the ceiling or the inner surface of the reflector and
c) on the walls of the room:

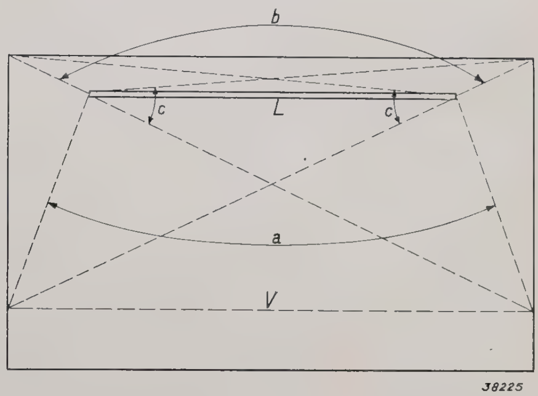

Fig. 1. The linear light source L is placed against the ceiling. The radiation a strikes the working surface V directly, while b is directed toward the ceiling and c toward the walls.

On the working surface, in addition to the direct radiation a), is also incident

1) part of the flux which has been reflected once by the ceiling or by the reflectors in which the lamps are placed,
2) part of the flux which has been reflected once by the walls of the room and
3) part of the flux which has been reflected repeatedly between the walls and the ceiling or between the various walls.

It is in general therefore permissible in such considerations to assume that only the part of the room which is higher than the working plane takes part in the repeated reflections, while the contribution from surfaces which are in or below the working surface may be neglected.

Of the four different contributions to the total illumination of the working plane which have been mentioned in the foregoing, the direct illumination will in general be by far the most important. This contribution should therefore be calculated accurately within reasonable limits. The other

[1]) Such tables are as a rule based upon the original investigations of W. Harrison and E. A. Anderson, Trans. I.E.S., 11, 67, 1916.

contributions are usually successively smaller in the order in which they are mentioned, and may therefore be calculated less accurately without causing important errors in the final result. Since, moreover, the calculation of these smaller contributions is fundamentally the same as in the case of ordinary lighting systems, or otherwise the same as that which will be given in this article for the direct illumination produced by the linear light sources, we shall here concern ourselves chiefly with the calculation of the direct illumination intensity.

The calculation of the direct illumination with linear sources of light

In this calculation of illumination provided by linear light sources we shall begin with the simple considerations which have already been given in this periodical [2]) about the light distribution of these sources. We assume that the surface of the TL lamp (*fig. 2*) radiates diffusely according to

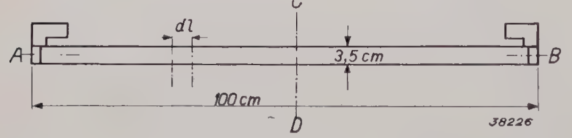

Fig. 2. Linear light source TL 100 with a length of 100 cm and a cross section of 3.5 cm. AB is the axis of the lamp and CD a perpendicular plane through the middle of the lamp, while a linear element of the lamp is indicated by dl.

Lambert's law. Observed from a sufficiently great distance the linear lamp then behaves as a point source would do which had a luminous flux distribution of the form indicated in *fig. 3*. Perpendicular to the axis of the lamp the maximum luminous flux of I_m c.p. is radiated, and in the direction of the axis of the lamp the light flux is zero, while in directions which make an angle of 90-α with the axis of the lamp the flux is given by:

$$I_\alpha = I_m \cos \alpha \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

As may be seen in fig. 3, the distribution of the luminous flux is now represented by a circle with a diameter I_m which is tangent to the axis of the lamp AB and revolved around it; *i.e.* the light intensity is a torus with no hole.

The light flux F (lumens) which is emitted by a point source of light with an isotropic (spherical) luminous flux distribution of I c.p. is given by:

$$F = 4\pi I.$$

The total luminous flux F which is emitted from a light source with a torus-shaped light intensity distribution according to fig. 3, becomes, however,

$$F = \pi^2 I_m.$$

For a linear light source with a length l one

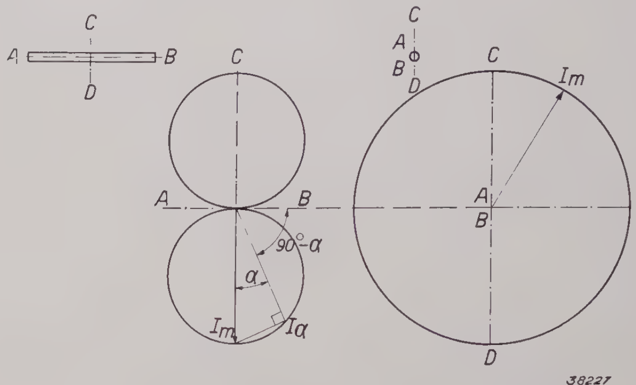

Fig. 3. Distribution of illumination from the linear light source AB in a plane through the axis of the lamp and in the perpendicular bisecting plane, respectively. I_m is the maximum luminous flux and I_α is the luminous flux at an angle 90°—α with the axis of the lamp AB. The spatial distribution of the luminous flux is obtained by revolving the lefthand figure around AB; it is thus a torus with no hole.

may speak of the luminous flux F_0 which is radiated per unit length, so that:

$$F = F_0 l$$

The linear lamp may thus be considered as a row of elements of unit length each having a maximum luminous intensity of

$$I_0 = \frac{I_m}{l} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

The luminous flux per unit length is then

$$F_0 = \pi^2 I_0 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (3)$$

while the total luminous flux radiated by a linear light source is given by

$$F = \pi^2 I_0 l.$$

On the basis of equations (1) and (2) it is now immediately clear that the contribution dI_α furnished by a cylindrical element of surface dl at the point P of a linear light source (*fig. 4*) to the luminous intensity in the direction PO making an angle of 90°-α with the axis of the lamp will be:

$$\mathrm{d}I_\alpha = I_0 \mathrm{d}l \cos \alpha \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

The contribution dE_α of this to the illumination (expressed in 10^4 lux, since we calculate with cm instead of m) at point O on the plane $ABCD$ situated at a distance h cm from the linear light source is given by:

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[2]) N. A. Halbertsma and G. P. Ittmann, Philips techn. Rev., 4, 181, 1939.

$$dE_a = \frac{dI_a}{r^2}\cos\gamma = \frac{I_0}{r^2}\,dl\cos a\cos\gamma,$$

where r represents the distance OP in cm and γ the angle between the direction of radiation PO and the vertical to the irradiated plane $ABCD$. If we call the perpendicular distance from the point O to the linear light source a (cm), then the contri-

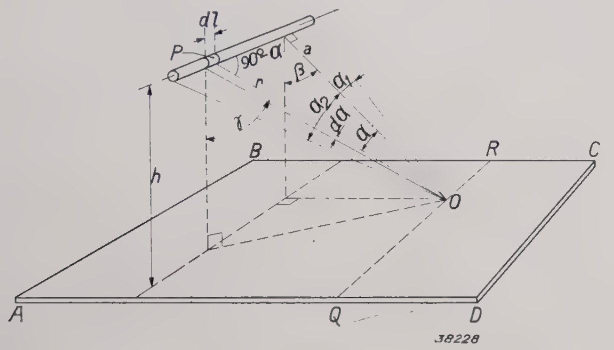


Fig. 4. Sketch showing the contribution to the illumination at the point O of the line QR on the working surface $ABCD$ which is furnished by the linear element dl at point P of the linear source.

bution to the illumination at O on the plane $ABCD$, furnished by a cylindrical element of the linear light source which is observed from O at an angle da, may be written as follows:

$$dE_a = I_0\frac{h}{a^2}\cos^2 a\, da \quad . \; . \; . \; . \quad (5)$$

The total illumination E which the linear light source produces at O on the plane $ABCD$ may now be found by integrating this expression (5) between the limits a_1 and a_2 keeping in mind that in fig. 4 the angle a_2 is positive and we must therefore take a_1 negative, because for the perpendicular a from O on the linear light source the angle $a = 0$. We then obtain the formula:

$$E = I_0\frac{h}{a^2}\cdot\frac{1}{4}\left|2a + \sin 2a\right|_{a_1}^{a_2}\cdot 10^4\text{ lux}, \quad . \quad (6)$$

where a must be expressed in radians, and h in cm, and I_0 represents the luminous intensity in c.p. per cm.

We shall consider this general formula (6) somewhat more closely for a simple case in which it will be found that it actually does give the familiar formula. If we consider the illumination E at distances a which are small compared with the total length l of the linear source, then

$a_1 = -\frac{\pi}{2}$ et $a_2 = +\frac{\pi}{2}$ so that we obtain

$$E = I_0\frac{h}{a^2}\frac{\pi}{2}\cdot 10^4\text{ lux} = \frac{\pi}{2}\frac{I_0}{a}\cos\beta\cdot 10^4\text{ lux},$$

where β represents the angle γ for $a = 0$. For a surface element which is perpendicular to the shortest line joining it with the linear source, $\beta = 0$, so that we then obtain simply that the illumination varies inversely proportionally with the distance:

$$E = \frac{\pi}{2}\cdot\frac{I_0}{a}\cdot 10^4\text{ lux} = 1{,}57\,\frac{I_0}{a}\cdot 10^4\text{ lux},$$

which is even more easily understood on the basis of simple considerations of symmetry, as given earlier in this periodical [3]).

Diagrams for the determination of the direct illumination

In practical cases in order to be able to find the direct illumination with linear light sources easily and rapidly, the most obvious method is to attempt to reduce the general formula (6) to a simple diagram.

In *fig.* 5 the plane is drawn through the linear source and the point O at which the illumination on the working surface is desired. It may immediately be seen that the total illumination E produced at O by the source of length l can be made up of the contribution E_2 of the light source l_2 and

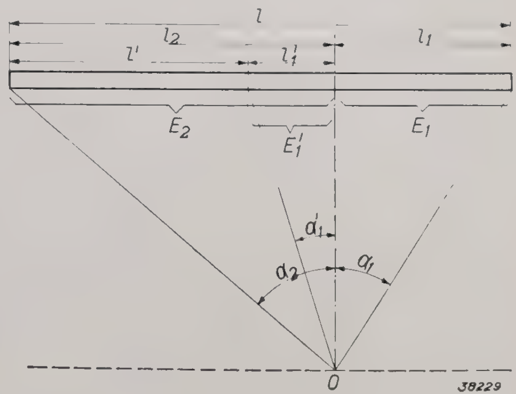


Fig. 5. The plane through the linear light source l and point O, at which the illumination on the working surface is to be determined, is here drawn to scale. The parts l_1, l_2 and l_1' of the linear source, which are seen from O at the angles a_1, a_2 and a_1', respectively, furnish the contributions E_1, E_2 and E_1', respectively, to the total illumination at point O on the working plane.

the contribution E_1 of l_1, while the illumination given by a light source l', on whose extension the base of the perpendicular from O lies, is of course the difference between the illuminations E_2 and E_1', which would be produced at O by the light sources l_2 and l_1' respectively.

This is all expressed (with correct sign) in formula (6) when we take a_2 and a_1' positive and a_1 negative, because the expression $^1/_4$ $(2a + \sin 2a)$ becomes zero for $a = 0$, so that, for example, the illumination E_2 given by the lamp l_2 at O is given by:

[3]) Philips techn. Rev. 4, 181, 1939.

$$E_2 = I_0 \cdot \frac{h}{a^2} \frac{2a_2 + \sin 2a_2}{4} \cdot 10^4 \text{ lux.} \quad . \quad (7)$$

This illumination E_2 at a point O of the line QR on the working surface $ABCD$ in fig. 4 can now be plotted in polar coordinates from O at an angle a_2 with the perpendicular to QR, as has been done in principle in *fig. 6*. In this diagram we have plotted

$$e = \frac{2a + \sin 2a}{4} \cdot 10^4. \quad . \quad . \quad . \quad . \quad (8)$$

In order to obtain the required illumination E_2, therefore, it is only necessary to multiply the value of e_2 at an angle of a_2 read off in fig. 6 by I_0/ha^2 [4]).

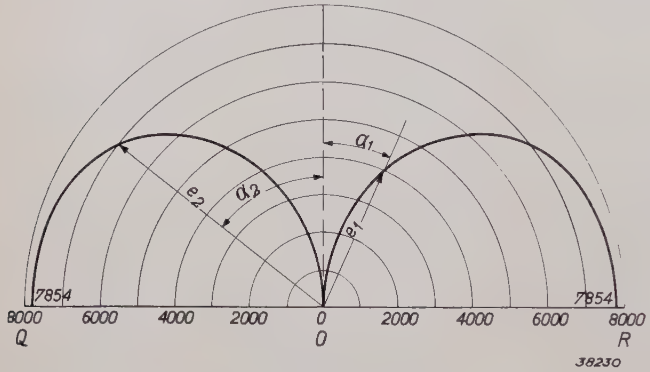


Fig. 6. The illumination E which a part of a linear light source with a luminous flux of I_3 c.p./cm produces at a point O which makes an angle a with the perpendicular distance a to the lamp amounts to I_3h/a^2 times the quantity e plotted in this polar diagram, when h is the distance from lamp to working surface in cm.

For the "Philora" lamps TL 100 which have a luminous flux F_0 of 10 lumens per cm, according to formula (3) we may assume about 1 for the candle power I_0 per cm. For "Philinea" lamps I_0 amounts to about 0.6.

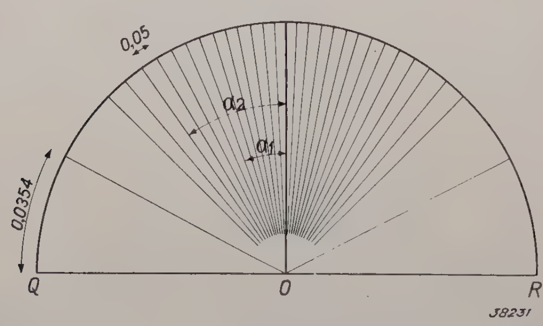


Fig. 7. In this polar diagram sectors are plotted each of which furnishes a contribution of 0.05 to the expression

$$\frac{1}{4} \left| 2a + \sin 2a \right|_{a_2}^{a_1}$$

A part of the lamp which, as seen from O, extends over n sectors furnishes a contribution to the illumination at O of 500 $n\, I_0h/a^2$ lux.

Usually the angles a_1 and a_2 will be so large that the corresponding values of e_1 and e_2 are easily read off if fig. 6 is simply laid over fig. 5 which is drawn to scale. For rays which pass through O at a small angle a to the perpendicular to QR, however, their point of intersection with the illumination diagram in fig. 6 is difficult to determine sufficiently accurately, and for these cases it is therefore advisable to use the diagram of *fig. 7*. This diagram is obtained by plotting sectors polarly from O each of which contributes the same amount to the direct illumination intensity on the working surface in the neigbourhood of O. For each sector of fig. 7.

$$\left| \frac{2a + \sin 2a}{4} \right|_{a_1}^{a_2} = 0{,}05 \quad . \quad . \quad . \quad . \quad (9)$$

so that between $a = 0$ and $\pi/2$, for which expression (9) becomes 0 and $\pi/4 = 0.7854$, respectively, lie fifteen complete sectors, while for the sixteenth and last sector expression (9) amounts to only 0.0354. If fig. 7 is sufficiently carefully drawn, the illumination at O can then be derived from it with a certain degree of accuracy, especially for fairly small angles a, in which region the successive sectors are almost equal in size, by laying fig. 5 drawn to scale over it. If, for example, the number of sectors in the diagram of fig. 7 compassed by the linear light source of fig. 5 seen from O is n, then according to formula (6) the direct illumination which it produces at O will be represented by:

$$E = 500\, n\, I_0 \frac{h}{a^2} \text{ lux} \quad . \quad . \quad . \quad . \quad . \quad (10)$$

Practical calculation of the distribution of the illumination intensity over the working surface

On the basis of a simple example we shall now explain somewhat more precisely how with the help of figs. 6 and 7 the direct illumination intensity can be determined at a sufficient number of points on the working surface to furnish data for a rough estimation of the illumination efficiency of a system of linear sources of light. We consider a room of 13 by 12 m with a height of 4 m. In this room 64 TL 100 lamps 1 m in length are placed against the ceiling in 8 parallel rows of 8 lamps each. These lamps are about 3 m above the working plane and at distances of 1.5 m from each other. On the working surface (*fig. 8*) we now draw the straight lines *II*, etc. up to *IX* vertically beneath the rows of lamps. It is along these lines that we wish to know the illumination intensity. At 1.5 m outside the lines *II* and *IX*, respectively, there are

[4]) In illuminating engineering the distances a and h are usually expressed in metres. The value of e in this polar diagram must then of course be decreased by a factor 100.

also drawn the lines *I* and *X*, both of which thus lie 25 cm outside the room. By allowing the distances between the lines to depend in a simple way upon those between the rows of lamps, provision may be

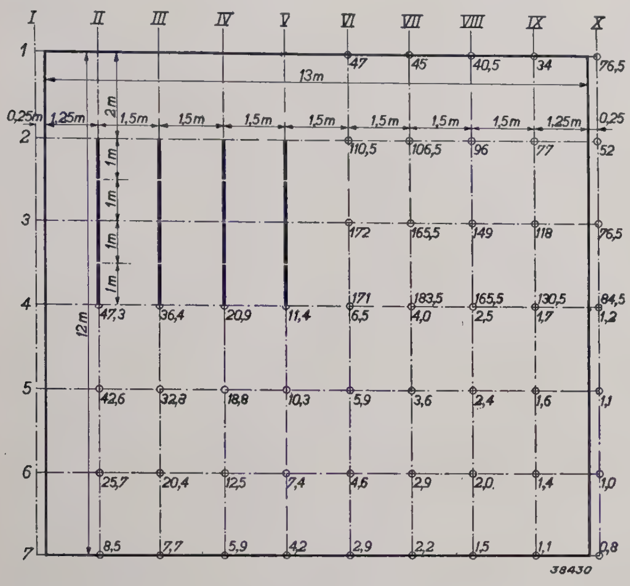


Fig. 8. In a room of 12 by 13 m and 4 m high there is a total of 8 times 8 linear lamps type TL 100 installed as indicated for the upper left-hand quadrant. In the lower half of the figure the contribution to the illumination of the row of lamps *II* is given for the intersection points of the working lines. In the upper right-hand quadrant the total illumination intensity is given for each of the points of intersection.

made that the successive lamps always give the same contribution to the illumination intensity at corresponding points on the successive lines. By a suitable choice of the working lines we thus simplify the calculation considerably.

We also draw several working lines in a direction perpendicular to the rows of lamps, and we have drawn them here 2 m apart so that there are seven of them (numbered from *1* to *7*) including the walls of the room. With the help of the diagrams of figs. 6 and 7 we now calculate for the points of intersection of these two systems of lines the contribution to the direct illumination furnished by every lamp. In the upper left-hand corner of fig. 8 the position of the lamps is indicated and in the lower half the contribution for every intersection is given which the row of lamps *II* furnishes to the total direct illumination of the working surface at that point. In the upper righthand corner of fig. 8 the direct illumination is noted for each of the intersections, rounded off to $^1/_2$, as furnished by the whole system at that point. The variation of this direct illumination in the other three quadrants is of course a mirror image of that in the quadrant calculated.

Along two mutually perpendicular lines *2* and *IX* in the upper right-hand quadrant we have now in *fig. 9* indicated the variation of the direct illumination. From this variation the points can immediately be found at which the direct illumination amounts to a whole number multiple of 10 lux. The mirror images of these points in the quadrant of the working surface lying to the left above in fig. 9 can thus be indicated without difficulty, and the lines may be drawn through them indicating equal direct illuminations, which give us the so-called isolux diagram for the upper left-hand quadrant of the working surface. Although the direct illumination in the middle of the room does not vary rapidly, even when the lamps are regularly spaced it falls off rapidly near the walls, and in the direction of the axes of the lamps it even falls to less than $^1/_3$ of its value in the middle of the room, as will be clear from fig. 9. If the greatest possible uniformity in light distibution over the whole working surface is desired, therefore, the light sources should be more closely spaced along the edges of the ceiling than has been done in this example.

If the luminous efficiency of a system is desired the total light flux which strikes the working surface should be determined, because the luminous efficiency is equal to this quantity divided by the luminous flux installed. From the illumination figures of fig. 8 by multiplication by the corresponding surface elements the total direct luminous flux which strikes the working surface can be relatively easily determined. For the quadrant under consideration with a surface of 39 cm^2 the direct luminous flux is 4 100 lumens, so that the average direct illumination will be about 105 lux. The ceiling and walls of the room to be illuminated have a reflectivity of 35 per cent. On the basis of the tables mentioned in the introduction (*cf.* footnote [1])), it can now be calculated that about 20 per cent of the light flux directed toward the ceiling finally reaches the working surface, while of the luminous flux directed toward the walls about 12 per cent also becomes effective. The total installed luminous flux is 64 000 lumens; of this nearly half is directed toward the ceiling, namely about 31 000 lumens. Of this half therefore 20 per cent finally reaches the working surface, *i.e.* about 6 200 lumens. Since the direct luminous flux to the whole working surface of 156 m^2 is $4 \times 4\,100 = 16\,400$ lumens and 31 000 lumens go to the ceiling, 16 600 lumens are directed toward the walls. Of this, 12 per cent finally reaches the working surface, *i.e.* 2 000 lumens. Due to the reflections at ceiling and walls, therefore, a luminous flux of $6\,200 + 2\,000 = 8\,200$ lumens reaches the

working surface of 156 m² and provides on it an increase of the average illumination of about 55 lux. This makes the average of the total illumination over the working surface $105 + 55 = 160$ lux. It must, however, be taken into account that after the passage of time the reflectivity of ceiling and walls will decrease, as well as the light flux given by the lamps. It is therefore advisable to count only on a value of 140 lux for the average illumination intensity. The total "effective" luminous flux which then strikes the working surface of 156 m² will now be about 22 000 lumens. Since the total light flux installed is 64 000 lumens, this means a luminous efficiency of nearly 35 per cent.

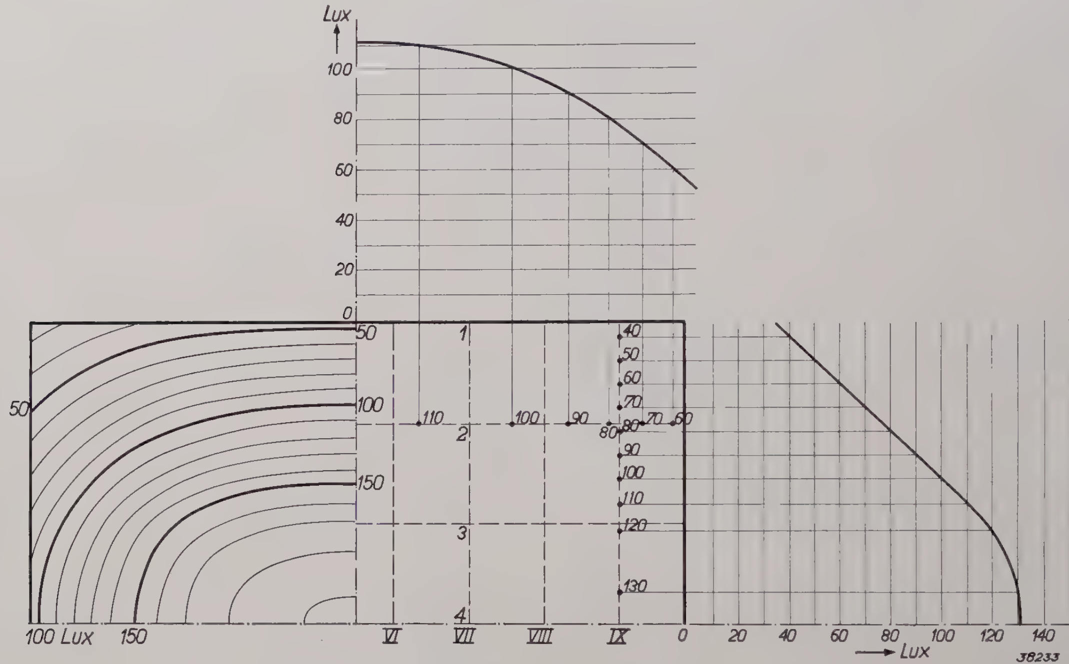
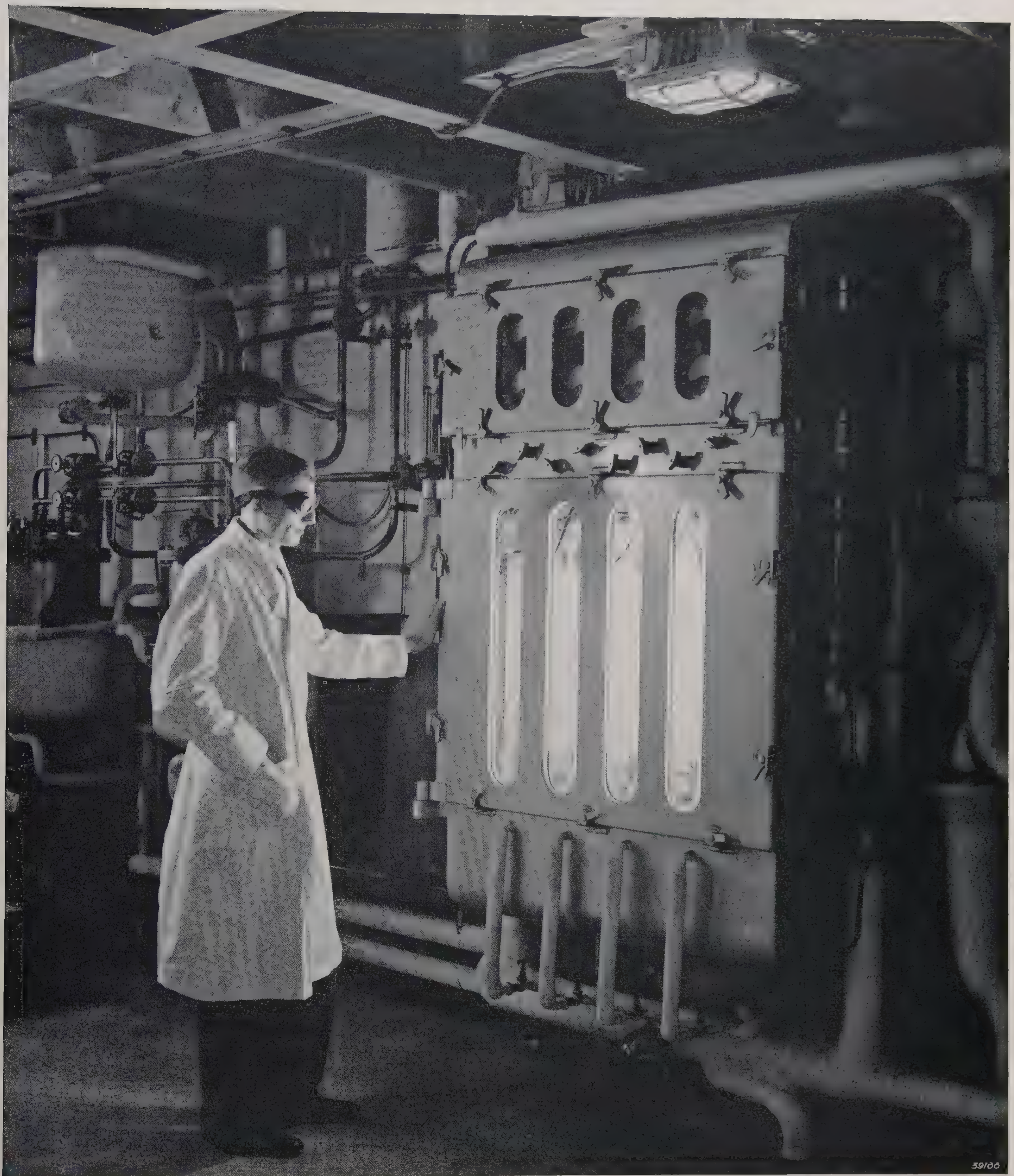


Fig. 9. For the working lines *2* and *IX* in the upper right-hand quadrant of fig. 8 the variation of the illumination is here drawn. The points along these working lines in which the illumination amounts to a whole number multiple of 10 lux are found graphically, and their mirror images can thus be drawn in the upper left-hand quadrant, for which we can in this way determine the isolux diagram of the direct illumination. In order to obtain a more uniform distribution of the illumination over the working surface it is found desirable to concentrate the lamps somewhat more toward the edges of the ceiling than was done in this example.

PREPARATION OF VITAMINE D



In the works of N.V. Philips-van Houten at Weesp the vitamine D is obtained from provitamine in this irradiating apparatus, which contains four high-pressure mercury lamps each consuming 1·5 kW.

A SIMPLE HIGH-FREQUENCY OSCILLATOR FOR THE TESTING OF RADIO RECEIVING SETS

by J. D. VEEGENS and M.K. DE VRIES.

621.396.615.1: 621.317.7

For the testing and adjustment of radio receiving sets after manufacture, as well as for purposes of service, a measuring oscillator has been designed which gives a high-frequency voltage whose amplitude is adjustable continuously between 1 μV and 0.1 V, while the frequency can be varied continuously between 100 000 c/s and 60 Mc/s (wave length 5 m). By means of a built-in low-frequency oscillator the amplitude of the oscillator signal can be modulated by 400 c/s. In this article the fundamental problems are discussed which are encountered in the construction of the measuring oscillator, particularly that of keeping the frequency stable, regulating the amplitude and preventing frequency modulation upon modulation of the amplitude of the oscillator signal. In conclusion several details of the construction of the apparatus are discussed.

For the testing and adjustment of radio sets an oscillator is needed which can deliver to the receiving set voltages, which have the same frequency and amplitude as the aerial voltages when the set is in use. In some cases a sinusoidal signal of constant amplitude will be enough, while in other cases the signal should be modulated in amplitude like the aerial voltages.

Upon closer examination of the possibilities of application of such an oscillator, it is found that the requirements which must be made of the apparatus may vary very much according to the nature of the application. For special measurements in laboratories and factories it may be necessary to know the frequency or the amplitude of the oscillator signal very accurately, or the form of the signal must be exactly that of a sine. In by far the majority of applications, however, less severe requirements suffice. In practice, precision measurements are not usually carried out with the oscillator, but only more or less qualitative tests. This is particularly the case for all tests which are carried out in service stations for detecting defects in radio sets. For such tests an oscillator signal of reasonably constant amplitude and frequency is also needed. The value of the frequency need not, however, be able to be adjusted with greater accuracy than about 1 per cent, while as to the amplitude, it is only necessary to know its order of magnitude.

In the following we shall describe a simple measuring oscillator (type GM 2 882) which is suitable for all applications where the highest requirements are not made. In order to lend the apparatus universal utility special care was taken that the frequency and the amplitude of the oscillator signal could be varied within very wide limits. Furthermore, emphasis was laid on practical features, such as easy operation and a strong and light construction of the apparatus. The measuring oscillator can easily be carried, which is of special importance for service purposes in order to be able to study faults in receiving installations on the spot.

Several fundamental problems

The problems which were encountered in the development of this apparatus may be divided into three groups:

1) Securing an oscillator frequency which can be varied within wide limits and which remains absolutely constant upon each adjustment of the measuring oscillator.
2) Securing an output voltage whose amplitude can be adjusted in a reproducible manner in a region with a lower limit of about 1 μV.
3) Securing the desired amplitude modulation of the signal.

The solutions of these three problems attained in our apparatus will now be dealt with successively.

Adjusting the frequency and keeping it constant

The frequency region of the measuring oscillator extends from 100 000 to 60 Mc/s, corresponding to wave lengths of 3 000 to 5 m. The lowest frequencies are intended for the investigation of the intermediate-frequency part of receiving sets, while the highest frequencies may be used for the investigation of the ultra short wave part of radio receivers. The adjustment of the frequency takes place in the usual manner by continuous regulation of a condenser and commutation of a self-induction coil. Six frequency regions may be chosen in this way, which are distributed as follows:

10— 30 kc/s
30—100 kc/s
100—300 kc/s
3— 10 mc/s
10— 30 mc/s
30— 60 mc/s

The principles of the connections of the oscillator are given in *fig. 1*. The frequency of the oscillations

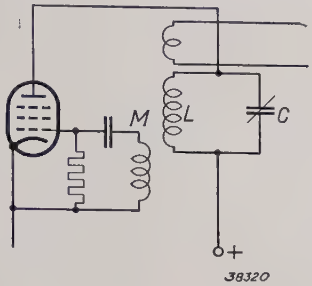


Fig. 1. Principle of the connection of the measuring oscillator. The oscillation circuit is not connected in the grid circuit, but in the anode circuit, since in this way a more nearly constant oscillator frequency is obtained.

excited is determined by the oscillation circuit LC according to the familiar equation:

$$f = \frac{1}{2\pi\sqrt{LC}} \quad \cdots\cdots\cdots \quad (1)$$

Equation (1) is actually only an approximation. There are numerous causes which result in a small displacement of the frequency, since the exact value of the frequency depends upon the resistances, capacities and self-inductions of all components which are coupled with the oscillation circuit, as well as upon the slope of the oscillator. These frequency shifts would of themselves be no objection if they could be taken into account upon calibrating the apparatus. Several of the quantities which affect the oscillator frequency, especially the loading resistance and certain capacities, exhibit, however, the undesired property that they may change with the conditions of performance. In order to prevent the occurrence of unallowable fluctuations in the oscillator frequency from these causes, very great attention had to be paid to these phenomena in the construction of the apparatus.

The influence of variations in the external loading was practically completely eliminated by not connecting the oscillator stage directly to the apparatus to be tested, but including an amplifier stage behind the oscillator stage. Variations in the working state of the amplifier stage will not react appreciably on the oscillator stage, since the two are only very loosely coupled. This will be dealt with later on.

We shall now go somewhat more deeply into the causes of capacity variations. The chief working conditions which may effect the capacities are the temperature of the surroundings and the supply voltage. The temperature of the surroundings affects mainly the dielectric constants of the insulation materials, whereby the capacities also experience a certain change. The thermal expansion of the tuning elements also results in a certain change in capacity, and, moreover, a change in the self-induction. It was, however, possible to make the temperature coefficients of the tuning elements so small that the temperature variations occurring in practice cause no unallowable fluctuations of the frequency. By suitable placing of the components, moreover, provision was made that the internal heating of the LC circuit during use is not too great. The result of these measures is that the frequency changes about 1 $^0/_{00}$ at the most due to the heating of the apparatus during use.

The influence of the supply voltage on the capacities is more serious. The valve capacities are not only dependent on the position of the electrodes in the valve, but also on the space charge between the electrodes. This space charge is in turn dependent on the anode current, which changes upon a variation of the supply voltage. Especially the capacity between cathode and control grid may hereby undergo considerable fluctuations which, without special precautions, would result in frequency variations.

The measures which may be taken to combat this phenomenon are of different kinds. The most conclusive means would be to stabilize the supply voltage. A suitable choice of connections for the oscillator valve may, however, help very much to keep the frequency variations mentioned small. We have used the following methods.

In the first place the influence of the valve capacity was in general limited by making the characteristic capacity C of the oscillation circuit as large as was possible in connection with the requirement that the oscillator must also be able to generate at short waves.

In order to make an oscillator generate it is necessary that the oscillation circuit should have a sufficiently high resonance resistance. The resonance resistance which can be obtained with an oscillation generally decreases with increasing capacity of the circuit. If therefore it is desired to obtain a sufficiently high resonance resistance at very high values of the resonance frequency, the capacity may not exceed a certain value. In order to make the value of the capacity at which oscillations can still be generated as high as possible, an oscillator valve with steep slope was used, namely the pentode EF 50[1]).

In the second place, as may be seen in fig. 1, the oscillation circuit was not placed in the grid circuit but in the anode circuit. Since the valve capacity between cathode and anode changes much less with the supply voltage than the capac-

[1]) The slope of the valve EF 50 is 6.5 mA/V. On the construction of this modern pentode see: A new principle of construction for radio valves, Philips techn. Rev. 4, 162, 1939.

ity between cathode and control grid, the frequency of the oscillation excited remains more constant with this arrangement. The influence of the changes of the grid-cathode capacity is not entirely eliminated in this way, since the capacity between control grid and cathode *via* the coupling coil M is transformed in the oscillation circuit in any case. In order to combat this effect the coupling between grid circuit and anode circuit is kept as loose as possible. Finally another simple measure is taken to combat the cause itself of the capacity variations, namely the fluctuations of the supply voltage. The rectifier for the supply is constructed for this purpose with a relatively high internal resistance. An increase in the supply voltage causes in general a fairly large increase in the plate current, and the internal voltage loss of the rectifier increases, so that a certain compensation of the voltage fluctuations is obtained. The result of all these measures is that the frequency variation with an oscillator frequency of 30 Mc/s for 10 per cent mains voltage change is not greater than about 6 000 c/s, *i.e.* 0.2 $^0/_{00}$. In the applications for which the apparatus is intended this variation may be entirely neglected. Any deviations between the frequency adjusted and that desired may then be ascribed more to the limitation of the accuracy of the scale than to frequency variations due to changes in the external conditions.

The regulation of the amplitude

In testing the sensitivity of a receiving set it is customary to apply to the aerial connections a high-frequency A.C. voltage such that a certain energy, 50 mW for instance, is fed to the loud speaker. The amplitude of the high-frequency signal then usually amounts to 1-50 μV in modern radio sets. The amplitude of the measuring oscillator must therefore also be able to be varied within this range. In the case of the apparatus under discussion, however, the voltage can be still further increased to 0.1 V. This is desirable when it is not a question of testing the set as a whole with the oscillator, but for instance only the intermediate-frequency stages which are designed for a considerably higher signal voltage.

The signal voltage given by the oscillator valve itself is of the order of magnitude of 10-100 V. If it is desired to make the oscillator generate stably, it is impossible to make the voltage much lower. In order to obtain the smallest desired signal, therefore, it is necessary to attenuate the voltage 10^7 to 10^8 times. An attenuation to 0.1 V is accomplished by coupling the "amplifier" stage following the oscillator stage only very loosely with the oscillator. By making this coupling loose enough not only is the desired reduction of the oscillator signal obtained, but at the same time, as already stated, there is the advantage that the influence of any changes in this amplifier stage or in the external loading is very much reduced.

Following the amplifier stage is an adjustable attenuator with which the signal voltage can be lowered continuously from 0.1 V to 1 μV (see *fig. 2*).

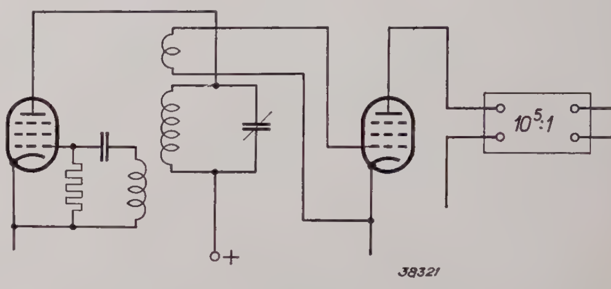


Fig. 2. Regulation of the amplitude of the oscillator voltage. The output voltage of the oscillator is applied to an amplifier stage *via* a very loose coupling. Following this is an attenuator which can regulate the voltage over a range 1 : 100 000.

In principle the attenuation of voltages is not difficult. An attenuator may be built up as a network of impedances, for which capacities, self-inductions or resistances may be chosen. In *fig. 3* an example is given of such an attenuator consisting of five cells, each of which attenuates the signal to $1/k$ of its original value. With $k = 10$, therefore, the desired attenuation by a factor of 100 000 will be obtained.

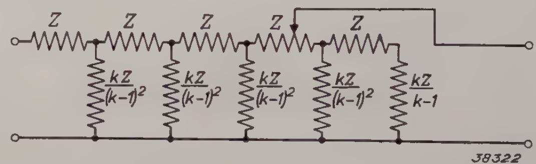


Fig. 3. Principle of an attenuator. In the network given each cell gives an attenuation of the input voltage by a factor k.

Since the attenuator must be able to function in a large frequency region, it is advantageous to use resistances as impedances, since resistances are in principle independent of the frequency. Practically, however, the branches of the attenuator will no longer behave exactly as resistances at the frequencies of short waves; the deviations become greater the higher the resistances. It is therefore advisable to choose small resistances. They must not, however, be so small that the self-induction of the necessary connection lines to the resistances becomes disturbing. A satisfactory compromise for the frequency region in question is formed by resistances between 10 and 300 ohms.

If the output signal must be very much attenuated by more than a factor 10^4 for example, all kinds of difficulties occur at high frequencies. In that case various components are situated in the neighbourhood of the output terminals whose potential fluctuates many thousand times as much as that of the output terminals themselves. If by means of capacitative or inductive effects a small part of these potential fluctuations is transmitted to the output terminals, an interference voltage is obtained at the output of the measuring oscillator which may easily be much greater than the desired signal.

In order to avoid the capacitative and inductive effects mentioned it is necessary to keep the connections short, to shield very carefully the components which generate high-frequency electric fields and to take care that the shielding is properly earthed. For some other parts of the connections also, particularly for the attenuator itself and for the supply part, proper earthing is of the greatest importance for the avoidance of interferences. To illustrate this fact we shall consider in more detail only a detail of the attenuator. As may be seen in fig. 3 the lowest connection line of all the transverse resistances, which is earthed, also serves as common connection terminal of the input and output voltage. Let us now assume that the earthing of the connection line between the transverse resistances takes place at some other spot than that of the common connection terminal, so that between the connection line and the connection terminal there is an impedance z, although very small. In *fig. 4* this case is required, the electrical diagram shows what happens in this case.

It may be seen from the diagram that an interference voltage E_s acts on the output terminals in parallel with the desired output voltage. E_s is given by the following:

$$E_s = E_i \, z/R_i,$$

where R_i is the input resistance of the attenuator, which has a value of about 300 ohms.

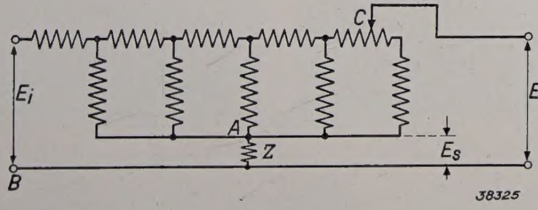


Fig. 4. Equivalent connections for the attenuator. Deviating from fig. 3, this diagram shows a certain impedance z between the connected transverse resistances of the attenuator and the corresponding connection terminal of the input and output voltage. This causes an interference voltage E_s which makes it difficult to obtain large attenuations.

If the input voltage must be able to be attenuated by a factor 10^5 without hindrance from this interference voltage, z/R_i must only be of the order of magnitude of 10^{-6}. z must therefore be smaller than 10^{-4} ohm. For a wave length of 5 m this means that the self-induction of the connection line must be smaller than 10^{-12} henry. A contact screw of 1 mm length may easily have a self-induction one hundred times as large. From this it will be sufficiently clear that in the construction of the attenuator very much care must be taken.

The impedance z in the attenuator, as already mentioned, is by no means the only source of interference voltages. Other interference voltages may be given, mainly by a capacitative coupling between different parts of the attenuator, furthermore by a capacitative coupling between the measuring oscillator and the apparatus to be tested, or by a coupling of these two apparatus *via* the power mains. Although all these effects can be combatted by the measures mentioned, their influence remains so strong that the attenuator may not be considered a precision instrument [2]). Since for certain applications, especially for selectivity measurement, it is advantageous to have at least one accurately known attenuation, a second attenuator which gives a fixed lowering of the amplitude in the ratio 1: 10 is introduced at the end of the connection cable which connects the measuring oscillator with the set being tested. This latter attenuator also serves as a socalled artificial aerial, *i.e.* it is built up of circuit elements of such kinds that the measuring oscillator seen from the output of the connection cable has about the same impedance as an ordinary aerial.

Securing the desired modulation

In sensitivity measurements on receiving sets the electrical energy which is delivered to the loud speaker is brought to a given value by regulation of the high-frequency voltage. Actually this energy is not determined by the amplitude, but by the modulation of the high-frequency signal, and therefore for a determination of the sensitivity it is necessary that the depth of modulation of the high-frequency signal should be known, while, moreover, it is desirable that the modulation should have a sinusoidal variation.

As to the frequency of the modulation, it is a

[2]) On the scale of the regulating knob of the attenuator voltage values are indicated. The actual output voltage may deviate from the voltage chosen as a result of the disturbing effects mentioned; the difference, however, amounts at the most to a factor 2. In this factor is included the drift of the output voltage with the frequency.

general custom to test the sets at a modulation frequency of 400 c/s. The reason for this is that this frequency is reproduced in full intensity by practically all sets, which is not the case for lower frequencies, while at higher frequencies some apparatus exhibit an irregular behaviour due to the occurrence of resonances. The modulation frequency of 400 c/s is generated by a separate oscillator included in the apparatus. Just as in the high-frequency oscillator, the resonance circuit is connected in the anode circuit. In order to obtain a truely sinusoidal signal inverse feedback is applied by means of a resistance in the cathode connection.

The A.C. voltage obtained in this way is now used to modulate the high-frequency signal, *i.e.* to influence it in such a way that there is a linear relation between the modulating voltage and the amplitude of the high-frequency signal. The method used here to accomplish this is so-called supressor grid modulation, whereby a negative bias voltage and in addition the modulating voltage are applied to the suppressor grid of a pentode. If there is a high-frequency A.C. voltage of constant amplitude on the control grid of a pentode, the amplitude of the anode A.C. is found to vary linearly with the modulating voltage within wide limits with a suitably chosen value of the suppressor grid bias voltage.

One of the problems which is encountered in the modulation of the oscillator signal is the occurrence of undesired frequency modulation. Suppressor grid modulation is accompanied by a periodic variation of the distribution of the space charge in the modulator valve (this is also true of every other method of modulation), and from this, as has already been mentioned, result periodic changes in the capacities between the electrodes. If the modulation were allowed to take place directly in the oscillator valve, which of itself is possible (see *fig.* 5), these capacity variations would lead to fluctuations of the oscillator frequency in the rhythm of the amplitude variations.

If such a signal which exhibits amplitude modulation as well as frequency modulation is now applied to the input of a selective high-frequency amplifier, the variation of the amplitude of the output signal may deviate very much from that of the input signal. The fluctuations of the frequency cause fluctuations in the output amplitude, especially when the frequency lies in the sloping part of the resonance curve. These amplitude variations are superposed on those which are caused by the amplitude modulation of the oscillator signal. The result may be a change in the depth of modulation, and at the same time a considerable distortion of the image of the modulation. *Fig. 6* gives two examples of this phenomenon: in case *a*) the

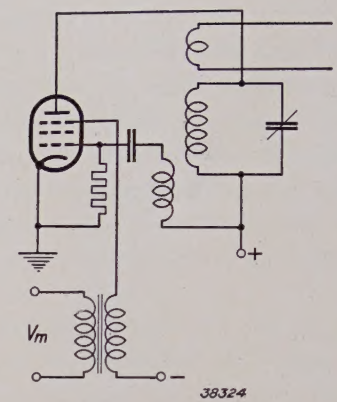
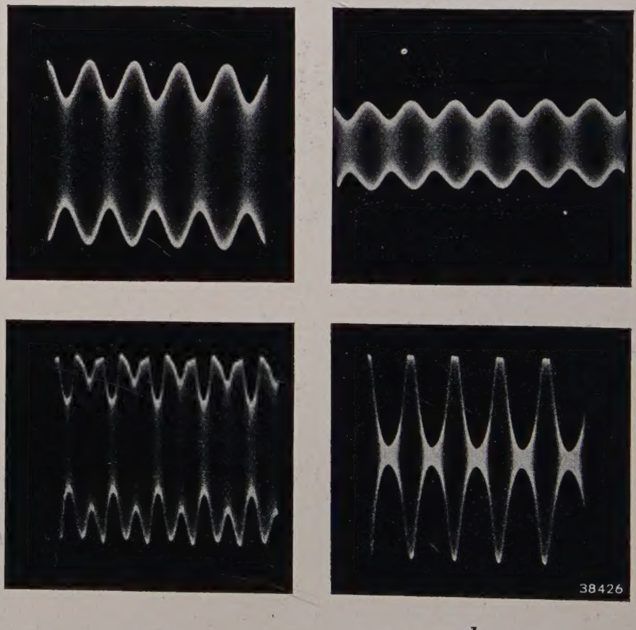


Fig. 5. Diagram for the modulation of the oscillator voltage with a signal of low frequency. These connections have the disadvantage that frequency modulation occurs.

average oscillator frequency corresponds to the resonance frequency of the receiving set, while in case *b*) the oscillator frequency lies on the sloping part of the resonance curve.

On the basis of the tuning curves of a large number of modern superheterodyne receivers a study was made of the degree of frequency modulation permissible when it is desired to be able to determine the sensitivity or the resonance curve of a



a *b*

Fig. 6. Examples of modulation distortion due to frequency modulation. A signal voltage which possesses amplitude modulation as well as frequency modulation is applied to an amplifier with a sharp resonance curve. In case *a*) the average frequency of the signal is equal to the resonance frequency, in case *b*) the signal frequency lies in the sloping part of the resonance curve. Above may be seen the input voltage, below the output voltage of the amplifier.

set with an accuracy of 10 per cent. It is found that for sensitivity measurements a frequency modulation greater than 200 c/s is not permissible, while for the measurement of the resonance curves of superheterodyne receivers this limit is only 30 c/s. But considering the fact that the selectivity of the receivers is determined practically only by the intermediate-frequency part, this last requirement only holds for relatively low frequencies.

How can the undesired frequency modulation be avoided? The modulation is made to take place not in the oscillator valve itself, but in the amplifier stage connected to it. Now the valve capacities of the amplifier valve do indeed change under the influence of the modulation, but since the amplifier stage, as already mentioned, is coupled only very loosely with the preceding oscillator stage, these variations have practically no influence on the resonance frequency of the oscillator circuit. In this way it is accomplished that the frequency modulation of 200 c/s permissible for sensitivity measurements is not exceeded in the whole range of oscillator frequencies up to 30 Mc/s. For frequencies up to 1.5 Mc/s (200 m) the frequency modulation even remains well within the limit of 30 c/s prescribed for selectivity measurements.

Construction of the apparatus

The larger the series in which an apparatus is manufactured, the greater the emphasis on easy methods of construction, while on the other hand the higher the permissible cost of the necessary tools. In this way in the manufacture of radio sets an extensive use of moulded parts has become prevalent. These parts necessitate expensive matrices, it is true, but they further simplify the manufacture very much.

Since the measuring oscillator is not manufactured in series of ten thousands, it was undesirable to make expensive matrices for the apparatus. For this reason no specially moulded parts were used. The chassis was built up as a closed body of sheet iron.

In the construction an attempt was made to collect the components as far as possible into small independent groups (units), which can be made and tested independently of each other. The wiring of the oscillator part was kept as short as possible, since the connection wires are always a source of individual differences between apparatus of a series, so that they might lead to difficulties in the calibration of the instruments.

A considerable shortening of the wiring of the oscillator circuit was obtained by a somewhat unusual construction of the wave length switch. The wave length switch is generally so constructed that it switches the connection terminals of the rotating condenser over to the different self-induction coils. For this purpose every coil must be connected to the wave length switch by means of a number of wires. In the measuring oscillator these connecting wires are avoided by using a rotating set of coils as wave length switch (see *fig.* 7). When the switch is given another turn, the

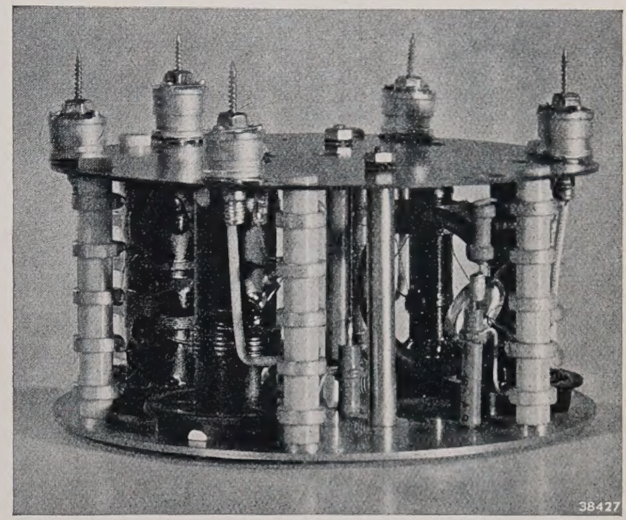


Fig. 7. The sets of coils of the measuring oscillator are mounted on the wave length switch so that upon switching over of the frequency region one set of coils (tuning coil and two coupling coils) is replaced by another set, while all connecting wires of the circuit remain in their places. For the region of shortest wave lengths an oscillator coil of one winding is used, which may clearly be seen in the figure; the remaining coils are wound on coil bobbins placed on the base plate.

self-induction coil and the corresponding back-coupling coil are replaced by another combination of coils, while otherwise all the connections and wiring remain unchanged. The capacities and self-inductions of the wiring thus also remain unchanged and may be kept very small, so that at every position of the rotating condenser one is certain that the wave lengths of the oscillator signals which are successively excited upon rotating the wave length switch have exactly the same relation as the square roots of the self-inductions of the coils. This has the advantage that for different frequency regions, such as

0.1— 0.3 Mc/s
1.0— 3.0 Mc/s
10 —30 Mc/s

the same scale can be used.

Fig. 8 shows the rear side of the apparatus opened, from which the arrangement of the main components may be seen. In order to avoid a heating up of the oscillator circuit as far as possible, the supply

part and all the valves are shielded with respect to the rotating condenser and the set of coils. Moreover, the latter are placed at the bottom, while the

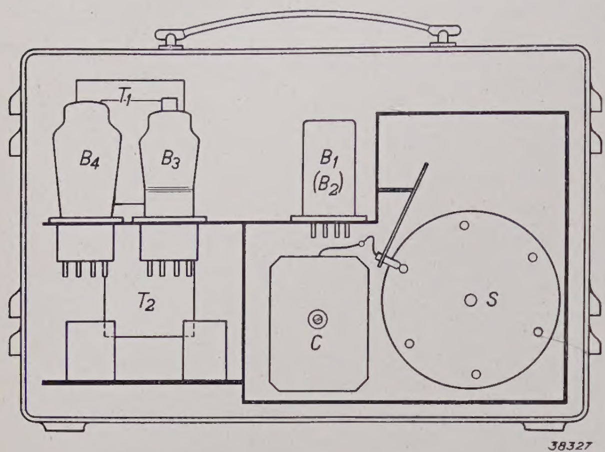


Fig. 8. Rear view of the opened measuring oscillator GM 2 882. B_1 oscillator valve shielded by a can, B_2 the valve directly behind B_1 of the "amplifier" stage, B_3 oscillator valve for the modulation frequency, B_4 rectifier valve of the supply apparatus; T_1 supply transformer, T_2 transformer of the modulator stage, C tuning condenser, S rotating set of coils. In the arrangement of the components the principle was followed that the components which develop the most heat should be placed at the top and separated from the condenser and set of coils by partitions.

sources of heat are at the top and can give off their heat by means of air circulation.

The rotating condenser is driven by means of a friction coupling *via* a second coupling which is flexible in an axial direction, so that external forces on the tuning knob can have no effect on the distance between the plates of the two sets. The pointer of the scale is driven by means of a string transmission by the condenser axis. By adjustment of the rotating condenser provision is made in every apparatus that the scale gives exactly the correct value at seven points: the possible errors at other points on the scale then remain within the limits of error in reading off. Scale and pointer are behind a glass window, so that damage during transportation is impossible. Together with the flexible transmission between the tuning knob and the condenser this offers the greatest guarantee that the calibration will remain accurate for a long time.

Fig. 9 finally gives a view of the outside of the measuring oscillator. In addition to the possibilities of adjustment already discussed, namely the switch for the wave ranges, the rotating condenser and the attenuator, there is only one other operating knob. This is a combination switch for switching the apparatus on and off and for passing from internal modulation to external modulation. In the latter case the desired modulation voltage can be applied from the outside; the valve of the low-frequency oscillator then serves as an amplifier valve.

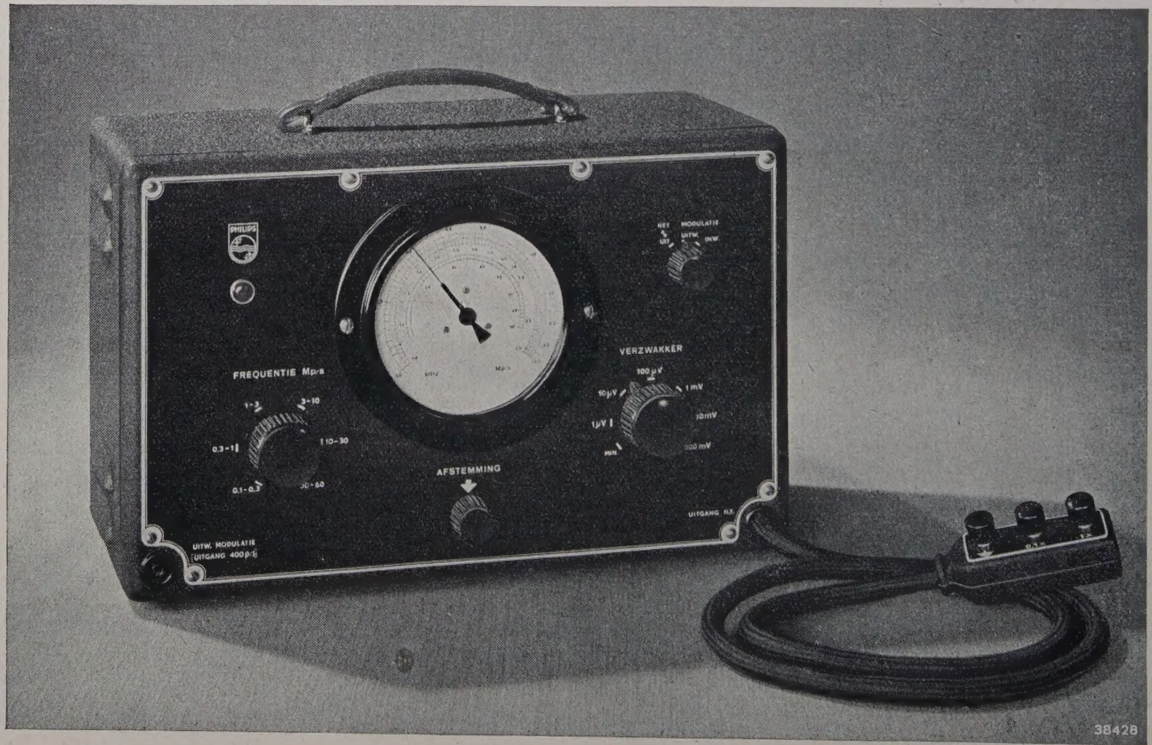


Fig. 9. View of the outside of the measuring oscillator GM 2 882. The apparatus has only four operating knobs, namely the tuning knob, the wave switch, the attenuator and a combination switch for switching the apparatus on and off and for choosing internal or external modulation.